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Oscillator Phase Noise Measurements using the Phase Lock Method

Master of Science Thesis

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ABSTRACT

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This thesis covers various methods to measure phase noise of oscillators. The main goal is to build a working phase noise measurement system for the Department of Electronics, Tampere University of Technology. The current facilities have limitations which restrict the measurements.

Phase noise of oscillators can be measured for example with the following methods: phase locked loop (PLL) method, delay line discriminator method, and cross correlation method. In addition to these basic methods, some commercial phase noise measurement systems will be covered. Phase noise of two-ports can be measured with a residual method, and this will be covered shortly.

The phase locked loop method is chosen for closer investigations. A measurement system based on that method is built and all related properties are introduced. Most of the required components were readily available, but two filters and a low noise amplifier for 0–100 kHz bandwidth was built and their performance verified during this project. The design and building process of those components is covered, and the performance of all the components used is verified with measurements. The measurement process and set-ups used are introduced.

The measurement system built during this project can be used to measure phase noise of oscillators, as shown in this thesis. Measuring phase noise of low noise oscillators is not an easy task, but it can be done without expensive commercial phase noise measurement systems using only commonly available equipment. This thesis shows how that can be done.

TIIVISTELMÄ

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Tässä työssä paneudutaan oskillaattorien vaihekohinaan ja sen mittaamiseen käytettyihin menetelmiin. Päällimmäisenä tarkoituksena on kehittää toimiva vaihekohinan mittausjärjestelmä Tampereen teknillisen yliopiston elektroniikan laitokselle. Laitoksen RF-Laboratoriossa tällä hetkellä käytetty menetelmä ei mahdollista useimpien oskillaattorien vaihekohinan mittaamista, koska järjestelmän rajat tulevat nopeasti vastaan.

Oskillaattorien vaihekohinaa voidaan mitata esimerkiksi seuraavilla menetelmillä: vaihelukkomenetelmä (PLL-menetelmä), viivelinjadiskriminaattorimenetelmä ja ristikorrelaatiomenetelmä. Näiden lisäksi käsitellään kaupallisia toteutuksia ja vaihekohinan mittaamista suoraan spektrianalysaattorilla. Kaksiporttien vaihekohina saadaan mitattua esimerkiksi residuaalimenetelmällä, joka esitellään lyhyesti.

Mittausmenetelmistä tarkimmin paneudutaan vaihelukkomenetelmään. Sen ympärille rakennetun mittausjärjestelmän ominaisuudet esitellään ja kerrotaan eri komponenttien vaatimuksista. Suurin osa järjestelmän vaatimista komponenteista löytyi laboratoriosta valmiina, mutta kaksi suodatinta sekä pienikohinainen vahvistin 0–100 kHz taajuuskaistalle täytyi rakentaa. Noiden komponenttien suunnittelu ja rakentaminen käydään läpi riittävällä tarkkuudella. Käytettyjen komponenttien ominaisuudet mitattiin niiltä osin kuin ne vaikuttavat käytettyyn mittausjärjestelmään. Näistä tuloksista ja mittauksista käydään läpi järjestelmän suorituskyvylle kriittiset alueet.

Projektin aikana rakennettu mittausjärjestelmä todettiin toimivaksi tavaksi mitata vaihekohinaa. Pienikohinaisten oskillaattorien vaihekohinan mittaaminen on varsin vaativa tehtävä, mutta tämän työn perusteella se on mahdollista tehdä ilman erityisiä kaupallisia vaihekohinamittalaitteita käyttäen hyväksi RF-laboratoriosta todennäkoisesti löytyviä laitteita.

FOREWORDS

This has been a really interesting journey. I have learned much and even though it has been quite hard once in a while, it has been kind of fun. First I would like to express my gratitude to the TUT Department of Electronics because they offered me this possibility. This was not a paid project, but nevertheless, it was still a good thing to have this possibility. Otherwise it would have been necessary to invent my own project, and that is never easy.

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Thanks for my whole family. Thanks dad for your encouragement and suggestions, even though the requirements and methods in the technical field are quite different than in your pedagogics field. It was still interesting to compare our simultaneous thesis writing projects and it also helped in some level. Thanks mom for your support and thoughts.

And finally, thanks for my lovely wife who has supported and loved me even though do not understand this field. Thanks also for your wonderful meals and thanks for our dogs who forced me to go outside so I would not lose my mind. You guys are awesome!

Tampere 27.5.2010

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CONTENTS

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TERMS AND SYMBOLS

Symbols

A	Amplitude of a signal
B	Equivalent Noise Bandwidth
c	Speed of light
E	Error
f	Frequency
f_0	Oscillator carrier frequency (Hertz)
f_m	Frequency offset from carrier (Hertz)
F	Device noise factor at operating power level A
G	Gain
H	A correction factor
k	Boltzmann's constant, $1.38 * 10^{-23}$ Joule/Kelvin
$\mathcal{L}_\mathrm{SSB}$	Single-sideband phase noise density (dBc/Hz)
l	Length
ϕ_0	Phase of a signal
P	Signal power
Q_L	loaded Q (dimensionless)
S	Shape factor of the filter
σ	Standard deviation
t	Time
T	Temperature
V	Signal
V_CC	Supply voltage
v_f	Velocity factor
V_tune	Tuning voltage of a VCO
ω_0	Angular frequency of a signal
U	Voltage
Z_0	Characteristic impedance

Terms

AC	Alternating current
AM	Amplitude modulation
ATT	Attenuator
BNC	Bayonet Neill-Concelman -connector
dBc	Level relative to the level of the carrier, in decibel scale
dBm	Power relative to 1 mW, in decibel scale, calculated as $10\log_{10}\left(\frac{\text{Power}}{1\text{ mW}}\right)$
DC	Direct current
DPDT	Double pole double throw -switch
DUT	Device under test
FET	Field effect transistor
FFT	Fast Fourier transform
FM	Frequency modulation
FR4	Flame resistant PCB material
GPIB	General purpose interface bus
IF	Intermediate frequency
LNA	Low noise amplifier
LO	Local oscillator
PCB	Printed circuit board
PLL	Phase locked loop
PM	Phase modulation
RBW	Resolution bandwidth
RMS	Root mean square
RSS	Root sum square
RF	Radio frequency
SMA	Subminiature version A -connector
SMD	Surface mount device
SSB	Single side band
USB	Universal serial bus
TUT	Tampere University of Technology
VCO	Voltage controlled oscillator

1. INTRODUCTION

Measuring physical phenomena accurately is both important for the science and difficult to do with high resolution. This thesis tries to give some information about one particular phenomenon, phase noise of oscillators, and how that can be measured with the required accuracy.

The first chapter covers the general things about this thesis. There will be some thoughts about the importance of this topic, what are the goals of this thesis and how this thesis is structured. The main motivation behind this thesis is to help practical every day measurement procedures in an RF laboratory and thus all theory and measurements will be presented with that in mind.

1.1 Motivation

Noise is everywhere. Sometimes it is wanted, usually not. Broadband noise is a good test signal when measuring for example the frequency response of an amplifier, because it includes all frequencies of interest. On the other hand, in communication systems noise is not a wanted phenomenon, because it disturbs the used signal and thus makes communication harder. It can thus be said that minimizing noise is a very important thing in every communication system. It is not possible to totally eliminate noise, but there are many ways to reduce it. When the used modulation schemes, circuits, and components are designed correctly, noise properties and noise immunity in a particular system can be significantly improved. During the design and evaluation process it is important to use the correct methods to measure all the necessary properties.

Measuring amplitude and frequency related properties is often a straightforward process. Phase noise, on the other hand, is more difficult to measure, still it is an important property when designing communication systems with good selectivity and sensitivity. Especially the phase noise of the oscillator(s) in receivers plays a very important role in selectivity.

A receiver adjacent channel selectivity, the total signal-to-noise ratio of a receiver and the velocity resolution of a Doppler radar are among things which are affected by single sideband (SSB) phase noise. [1] Without a method to measure phase noise accurately, it is impossible to understand all limitations in the communication system. This understanding is important when a designer tries to improve the communication system.

1.2 Goals of the Thesis

The main goal of this thesis is to build a phase noise measurement system for the radio frequency (RF) laboratory of the Department of Electronics, Tampere University of Technology (RF-Laboratory in the following sections). Different phase noise measurement systems are also investigated during this project even though only one is built and its performance verified. The emphasis is on radio frequency oscillator measurements, but naturally most of the procedures described in this thesis can be adapted to measuring other oscillators as well. After reading this thesis, the reader should be familiar with phase noise measurements and related problems which will arise when building a suitable measurement set-up.

It was known in advance that the current phase noise measurement system in the RF-Laboratory was not adequate. At the moment phase noise is usually measured directly with a spectrum analyzer, and this method is a limited one. Basically it is possible to measure only oscillators in a limited performance range, because the system used sets some limitations for the performance of the oscillator. This method can not be used to measure oscillators with good phase noise performance, because the noise floor of the measurement is too high. Also oscillators which drift much are problematic, because the carrier frequency should be steady during the measurement. This particular method will be covered and the problems discussed in more detail in sections 3.5.1, 3.5.1, and 5.2.1.

At the beginning of this project it was planned that a working measurement set-up would result from this project. During the project that goal changed somewhat, and finally the goal was to acquire as much information of phase noise measurements as possible. If some working solution could be found, necessary components would be ordered or built and the measurement set-up assembled. Finally the performance of the assembled measurement set-up would be verified thoroughly.

1.3 Structure of the Thesis

Chapter 2 describes the theory behind phase noise. The discussion is limited to the parts necessary for this thesis. References to more detailed sources are given.

Chapter 3 reviews the different measurement set-ups and also introduces the problems found in those set-ups. It is necessary to know the weaknesses and strengths of different methods to choose a suitable one.

Chapter 4 describes in more details the chosen measurement set-up, which is called the PLL method. Different parts of the set-up are introduced and important properties discussed. This chapter covers also some specific things which must be considered when building this or any other measurement system.

Chapter 5 shows the measurement results and provides more information on how the actual measurements were done. The chapter begins by describing all necessary preparative measurements and continues finally to the results from the actual phase noise measurements. Problems arising during the different phases of the measurement process will be covered and solutions suggested, if possible. One section is also reserved to error analysis.

Chapter 6 discusses on the biggest weaknesses of the chosen measurement set-up and gives some ideas how to improve the performance. There are some areas where more research is needed, and that information can be found in this chapter. This chapter also summarizes the whole project.

2. WHAT IS PHASE NOISE?

Phase noise is closely related to the performance of oscillators. Naturally the phase noise of a whole complex system can be measured, but it can be closely approximated by adding the phase noises of different oscillators together. In this context it is sufficient to define phase noise and all the related properties through the behavior of an oscillator.

The output of an ideal oscillator can be expressed as

$$V_{\text{out}}(t) = A_{\text{out}} \sin(\omega_0 t + \phi_0) \tag{2.1}$$

where A_{out} is the amplitude, ω_0 the angular frequency and ϕ_0 the phase angle of the signal. Eq. (2.1) describes a theoretical situation because it does not take any kind of noise into account. When also noise is considered, the amplitude and the phase angle become dependent of time and the equation can be written as: [2, 3]

$$V_{\text{out}}(t) = A_{\text{out}}(t) \sin[\omega_0 t + \phi(t)] \tag{2.2}$$

In a well designed and behaving oscillator the amplitude is so stable that it can usually be considered constant. Usually the noise level caused by the amplitude modulation component ($A_{\text{out}}(t)$ in the equation) is 20 dB lower than the noise level caused by the phase noise component ($\phi(t)$ in the equation), because most oscillators operate in saturation. Thus phase noise is the main noise contributor in an oscillator. [2]

The phase can vary either discretely or continuously. The previous one produces discrete signals called spurious frequencies, which can be seen as discrete components in the spectral density plot. The continuous part of the phase fluctuation causes the phenomenon called phase noise. It is a random phenomenon and thus can only be described with statistical terms. It produces sidebands to the spectral density plot which can be observed for example with a spectrum analyzer. Figure 2.1 show an example plot which includes phase noise and some spurious signals. [2]

Phase noise has been studied much in the past, and there are many equations covering its behavior. Probably the most used one these days is the Leeson's equation [1, p. 230]

$$\mathcal{L}_{SSB} = 10\log_{10}\left[\frac{FkT}{P}\frac{1}{8Q_{\text{L}}^2}\left(\frac{f_0}{f_{\text{m}}}\right)^2\right] \tag{2.3}$$

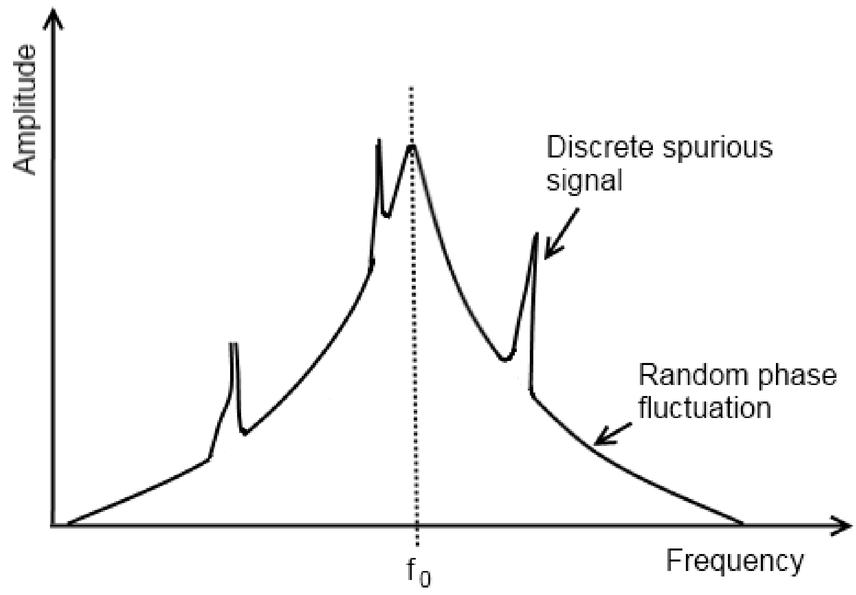


Figure 2.1: A general phase noise spectrum

where

$$
\begin{aligned}
\mathcal{L}_{SSB} &= \text{single-sideband phase noise density (dBc/Hz)} \\
F &= \text{device noise factor at operating power level } A \text{ (dimensionless)} \\
k &= \text{Boltzmann's constant, } 1.38 \times 10^{-23}\ \text{J/K} \\
T &= \text{Temperature (K)} \\
P &= \text{oscillator output power (W)} \\
Q_{\mathrm{L}} &= \text{loaded } Q \text{ (dimensionless)} \\
f_0 &= \text{oscillator carrier frequency (Hz)} \\
f_{\mathrm{m}} &= \text{frequency offset from carrier (Hz)}
\end{aligned}
$$

Eq. (2.3) applies when f_m is greater than the $f_1 = 1/f$ -flicker corner frequency of the active device and smaller than $f_2 = f_0/2Q_L$. In (2.3), f_1 is typically less than 1 kHz and f_2 is the frequency where white noise level starts to dominate, usually some MHz. This is shown more clearly in Figure 2.2. There are also some other restrictions which must be met before the equation can be used: F must be known; loaded Q must include all effects from component losses, and device and output buffer loading; and the oscillator must include only a single resonator. [1]

Leeson's equation identifies the different parameters affecting the phase noise performance in an oscillator circuit. It can also be easily seen what parameters must be tuned and how if we want to reduce phase noise. Some of them can be easily measured or de-

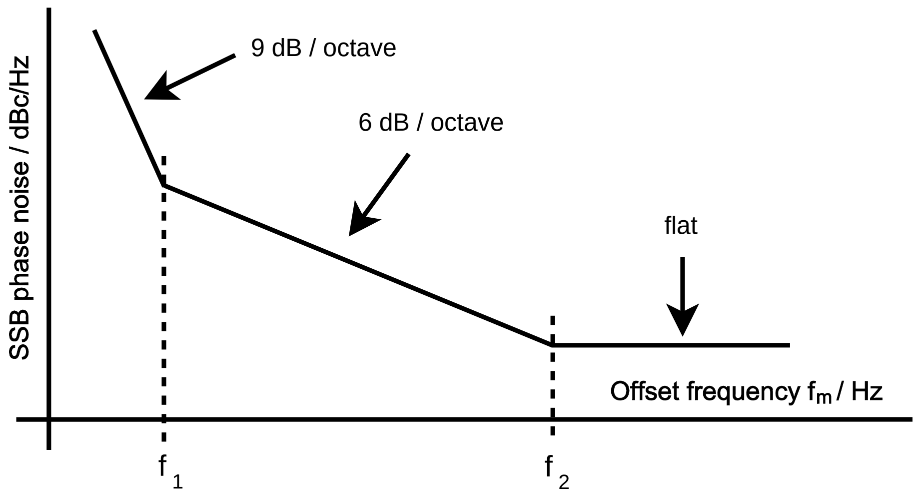


Figure 2.2: The different areas of a phase noise frequency plot

termined, but some are more problematic. The most complicated one is the noise factor of the oscillator (F), because of the non-linear operation of the active devices in the oscillator circuit. In practice F is often not measured directly. Loaded Q is obtained with a suitable technique, and after measuring SSB phase noise, F is calculated from Leeson's equation. [1] So, even though Leeson's equation can be used to calculate the phase noise of an oscillator, it is usually quite difficult to use it for that, but during the design process it can be used as a guide and a checking tool.

Phase noise is usually expressed as dBc/Hz at some specific offset from the carrier. That means noise level relative to the carrier level and calculated in 1 Hz bandwidth. Often only single sideband noise is considered, because the noise sidebands are assumed to be approximately symmetric around the carrier. Many measurement set-ups measure both sidebands and the final result is an average between the sidebands.

Phase noise describes how the frequency of an oscillator varies in short time scale. The long term frequency stability is called frequency drift, and it must also be considered during the measurement process. The output frequency of an oscillator takes some time to stabilize after the oscillator has been started and this drift can be up to dozens of MHz as was noticed during this project. The output frequency also usually drifts noticeably during the measurements, especially in the case of free running oscillators. This drift is a real problem, because during the measurements the system must be able to lock to the carrier or the carrier must be stable enough. If a carrier tracking mechanism is used, the drift is not such a big problem anymore.

3. MEASURING PHASE NOISE

Usually there are more than one way to measure the phenomenon of interest. This is a good thing, because all measurement systems have strengths and weaknesses. The engineer must understand the limitation so a suitable method can be chosen.

This chapter covers the different phase noise measurement methods and some space is also reserved for the discussion about the general things related to all measurements. The phase noise measurement methods covered in this thesis are *PLL method*, *Delay line discriminator method*, *Cross-correlation method*, and *Commercial products*. The first one is the simplest and has the biggest limitations, and the second last one requires the most complex measurement system, but is also the most versatile one. The *Cross-correlation* method can even measure oscillators with a better phase noise performance than the reference oscillator.

3.1 About Phase Noise Measurements

Ray Kammer said when he was the director of National Institute of Standards and Technology: 'If I have one watch, I always know what time it is. If I have two watches, I'm never quite sure what time it is'. [4] This is the basic problem in all measurements: getting reliable and reproducible results is difficult. Especially when measuring some non-static property, such as a real world signal, deviations in results are unavoidable. These deviations depend mostly on the measurement accuracy. If 1.5 GHz is accurate enough frequency for an oscillator, it is usually possible to get that accurate result all the time, but if the required accuracy is 1.500 000 000 GHz, some deviations will be seen in the results. The measurement resolution is thus one of the things which must be considered when choosing a suitable measurement system.

In general, measuring phase noise is more difficult than measuring amplitude or frequency related properties. In reality this of course depends on the required accuracy. The biggest reason for this is the huge dynamic range required in the phase noise measurement process. Good oscillators can have phase noise in the range of -120 dBc/Hz at a 10 kHz offset from the carrier. Another limiting property is the phase noise of the measurement instrument, which can easily prevent using some of the techniques described in the following sections.

There are many different phase noise measurement techniques, and the right one must be chosen when performing measurements. It is necessary to know the weaknesses and strengths of different systems, because none of these methods is a perfect choice for every situation. Often the most limiting part is still the availability of measurement systems, because specialized phase noise measurement equipment is relatively expensive. Fortunately there are some in-expensive systems, and one of them is built and its performance investigated during this project. Section 3.2 describes in more details the chosen method, which is called the *PLL method* in this thesis.

There are many ways to categorize phase noise measurement systems. One way is to divide them based on how the signal is measured, into direct method and in-direct methods with some form of a conversion. In the first ones phase noise is measured directly from the signal and in the latter ones the signal is somehow converted to a form which is more suitable for measurement purposes. Usually the in-direct method does not directly measure the phase noise, but some other property, for example the error signal in a PLL loop, is measured and phase noise is calculated from the results.

3.2 PLL Method

Because many measurements are difficult to make in high frequencies and much easier near 0 Hz, it is often easier to bring the output of the DUT to near 0 Hz, if that will not alter the interesting property too much. Every traditional radio transmitter and receiver makes this conversion, so components for that conversion can be found easily. Naturally quality is a much more important feature in measurement systems than for example in consumer radio receivers, so care must be taken when choosing the right way to do this conversion.

In [5, p. 9.10] a low cost method for measuring phase noise is introduced. The measurement set-up can be seen in Figure 3.1. In that method there is a reference oscillator, which should have as low phase noise as possible. The phase noise of this oscillator is the most limiting factor. The signal of the reference oscillator is mixed with the signal of the DUT and the reference oscillator is kept locked to the DUT with a phase locked loop (PLL).

PLL is basically a system that compares the phases of two input signals and generates a third signal which can be used to steer one of the input signals closer to another. When the phases of the input signals are aligned, the loop is said to be locked and the control signal is zero. The mixer behaves as a phase detector and its output is zero when the input signals are in quadrature. References [6] and [7] describe in detail why a mixer can be used as a phase detector. In this thesis, the varying DC level is the control signal which is used for locking the PLL.

When the frequencies of the input signals are close enough to each other, the output of the double-balanced diode mixer is basically a DC voltage which relates to the phase difference between the input signals. This voltage varies a little due to phase noise of the input signals. It should be remembered that the noise present at the output of the mixer

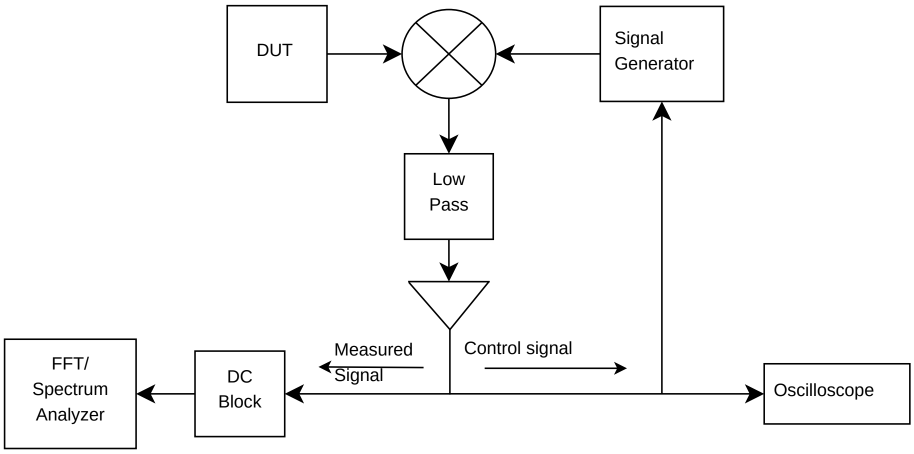


Figure 3.1: PLL method set-up

includes phase noises of both oscillators. If the noise from the reference oscillator is more than 20 dB lower than the noise from the DUT, the main contributor for phase noise is the DUT, as Table 3.1 shows. This topic will be discussed later in this section. The little voltage variations in the PLL control voltage are amplified and measured with a spectrum analyzer. The method for measuring the frequency content of the signal is not critical. For example an FFT or a traditional RF spectrum analyzer can be used. The exact DC level of the signal is not interesting, but the interesting part is the variation in the control voltage. The DC level is thus filtered away with a DC block. [6], [5, p. 9.12]

This method is relatively easy to deploy and can be used to measure phase noise down to 170 dBc/Hz at 10 kHz offset, as Figure 3.2 shows. It also shows noise floors of two other measurement techniques, which will be described in the next sections. [8]

The PLL method is also suitable for very broadband measurements. With only a few different mixers and suitable reference oscillators, frequencies from 1 MHz to dozens of GHz can be measured. The main problem with this method is that it is not possible to know which part of the noise comes from the reference and which from the DUT, but this is the problem in most measurement systems. If the used reference is good enough, so that its noise is 20 dB lower than the noise of DUT, it can be neglected. If the noise levels are not that far from each other, a correction factor must be subtracted from the result. The correction factor is from 0 to 3 dB, where the highest number is used when the noise levels are equal. [8]

Table 3.1 shows the correction factors for different noise level differences. They can be calculated as

$$P_{\text{corr}} = 10\log_{10}\left(1 + 10^{\frac{-\Delta P}{10}}\right) \text{dB} \tag{3.1}$$

where P_{corr} is the number that must be subtracted from the results, and ΔP is the difference between the noises of the reference and the DUT in dB. [9]

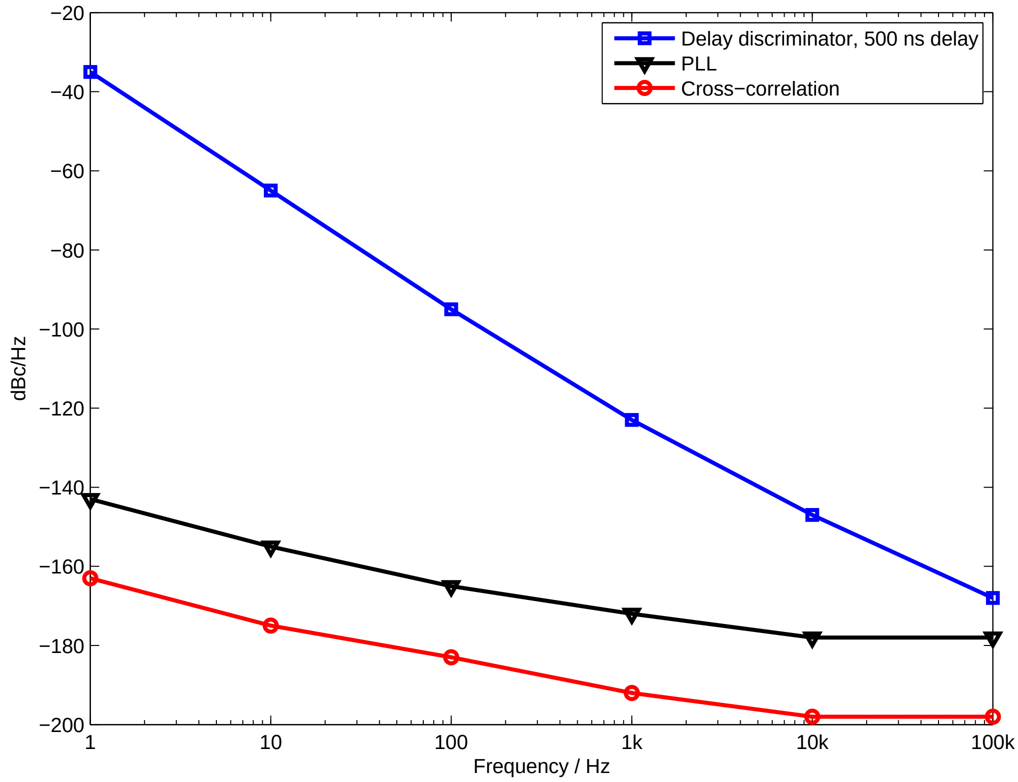


Figure 3.2: The expected phase noise measurement floor

Table 3.1: The correction factors if the phase noise of the reference oscillator is near the phase noise of the DUT

$\Delta P/\mathrm{dB}$	0	2	4	6	8	10	15	20
$P_{\mathrm{corr}}/\mathrm{dB}$	3	2.12	1.46	0.97	0.64	0.4	0.14	0.04

If the DUT is a VCO this method can be altered a little. The reference oscillator will then be a free running one, and the DUT would be controlled with the PLL. We just need a suitable PLL amplifier after the low pass filter. This variation and a suitable amplifier is described in [10] and [11]. The altered version can measure only VCOs, which must be considered when choosing the measurement set-up.

The method described in this section was used in this thesis even though the real measurement set-up is more complex. The configuration of the used measurement system is described in more details in Chapter 4 and its performance is verified in Chapter 5.

3.3 Delay Line Discriminator Method

Like the PLL method, the delay line discriminator method utilizes a mixer, but the methods are otherwise not similar, as can be seen in Figure 3.3. For example no reference oscillator is needed. In this method the output of the DUT is amplified and then ran through a 10 dB directional coupler. The through path of the coupler is connected to a delay line,

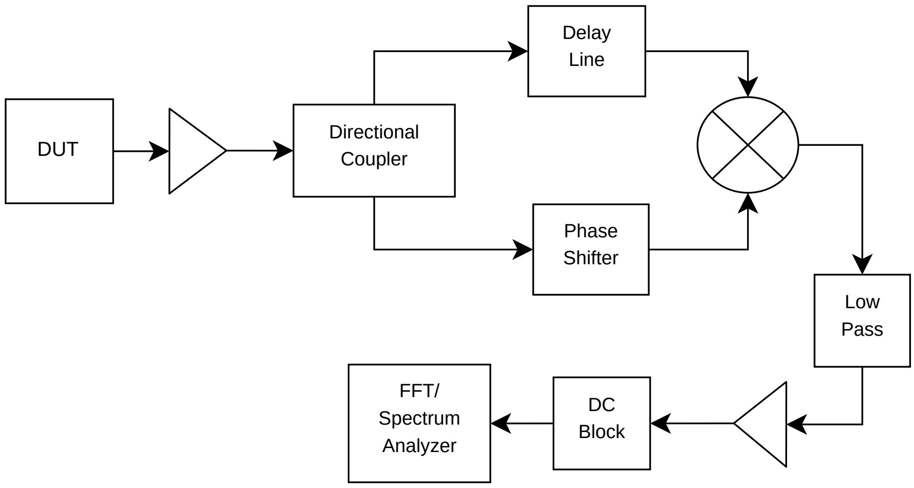


Figure 3.3: Delay line discriminator set-up

for example 100 m of cable, and then to the mixer input. The 100 m cable has been used in [8], because it provides around 480 ns of delay, as Eq. (3.2) shows, and usually around 10 dB of attenuation. Thus the amplitude of the delayed signal is about the same as the amplitude of the coupled signal which was attenuated 10 dB in the directional coupler. [8] If a delay line without any additional attenuation is used, the directional coupler can be changed to a power splitter.

The following equations shows how the delay from the specific cable can be calculated. The equation offers a theoretical value for the delay, which should be close to the real one, but it is not exactly the same. Connectors add delay and the cable properties vary inside the cable.

$$t_{\text{delay}} = \frac{l_{\text{cable}}}{v_{\text{f}} c} = \frac{100\text{ m}}{0.7c} \approx 480\text{ ns} \tag{3.2}$$

$$v_{\text{f}} = 1/\sqrt{\varepsilon_{\text{r}}} \tag{3.3}$$

where v_{f} is the velocity factor and ε_{r} is the relative dielectric constant of the dielectric in a coaxial cable. ε_{r} is usually from 1.3 to 2.3 and thus v_{f} is in the range of 0.66 to 0.88 [12].

The coupled signal is connected to a phase shifter and then to the other input of the mixer. The phase shifter is used to adjust the mixer inputs to quadrature. Again the output of the mixer is measured. This time the output is not directly the phase noise, but it must be converted to the right form. When the mixer sensitivity is calibrated correctly, the results must be divided by f^2 to get phase noise level, where f is the used offset frequency. The mixer sensitivity can be obtained by shifting the carrier frequency up and down from the center frequency to get approximately ± 1 V change. [8]

The good thing with this method is that it can be used to measure noisy sources. On the other hand, it does not work with good sources, because the noise performance of this method is the limiting factor, as Figure 3.2 shows. The noise floor depends on the length

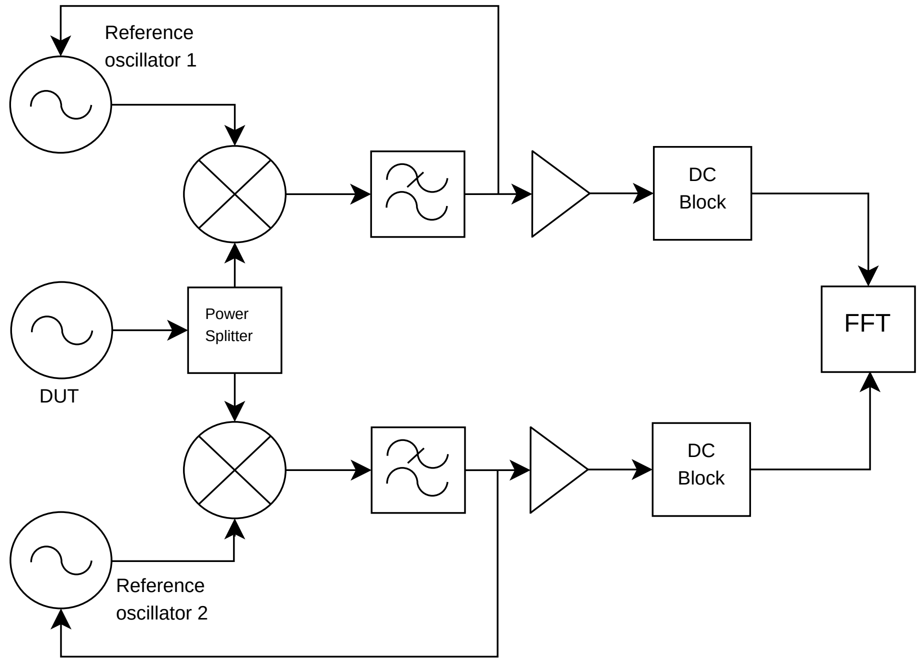


Figure 3.4: Cross correlation method set-up

of the delay. The longer the delay the lower the noise floor, but it will also mean higher losses and lower offset frequency. [8]

The highest usable offset frequency depends mostly on the length of the delay. There is a null at $f = \frac{1}{t_{\text{delay}}}$ offset frequency, and the recommendation is to use offset frequencies up to $f = \frac{1}{4t_{\text{delay}}}$. With a 500 ns delay, the usable offset frequency range is from 0 to 500 kHz. [8]

3.4 Cross-Correlation Method

The cross-correlation method is built around a similar measurement set-up as the PLL method, described in Section 3.2, although it is more complex. As can be seen in Figure 3.4, there are two reference oscillators, one power splitter, two mixer/amplifier/PLL circuits and a cross-correlation FFT analyzer. The output of the DUT is connected through a power splitter to two mixer circuits where it is mixed with the signal from those two reference oscillators. The outputs of the mixer circuits are used for PLL circuits as in the PLL method and these PLL signals are again of interest. The mixer output signals are then amplified, the DC is filtered away and finally the signals are fed to two channels of the FFT analyzer. The FFT must be able to perform a cross correlation measurement between the two input signals. This results in preserving the common noise, whereas the noise which is not common in both signals is attenuated. The common noise is the noise from the DUT because it is the same in both signals and the noise which is not common comes from the two reference oscillators. [8]

This method offers 15 to 20 dB better noise performance than the PLL method, so it can be used to measure oscillators with lower phase noise. It is even possible to measure oscillators with better noise performance than the reference oscillators, because the noises from the reference oscillators are reduced significantly. [8]

On the downside, wit the cross-correlation method, many measurements must be made and the average calculated between them. Thus the measurement takes longer, and the DUT must be kept locked longer time. One sweep takes approximately 10 s, and the required amount of sweep is 2^n where $n > 2$. With a noisy source this may not be easy. [8] During the measurement phase of this project, it was noticed that even three sweeps was too much when using the PLL method if the DUT was a free running oscillator. The frequency of the DUT drifted so much that the PLL went out of lock during the measurement. By carefully tuning the frequency of the reference oscillator the lock could be preserved. This is probably a limitation in the used PLL scheme, but no alternative methods were available for testing.

The cross-correlation method is thus suitable only for measuring good oscillators with small drift. It also requires more equipment when compared to the PLL method, though the required components are easy to find. Often there is a power splitter available in an RF laboratory, and suitable mixers are not very expensive. For example Mini-Circuits ZX05-42MH-S+ costs around 30 € at the time of writing this thesis and probably the price is not a problem. In this thesis the cross-correlation method was not used, but in the PLL method the mixer was ZX05-42MH. It is the same model as ZX05-42MH-S+, except a bit newer generation. The mixers and low noise amplifiers should be the same model, so that their performance would be similar.

3.5 Commercial Measurement Systems

Many companies manufacture specialized phase noise measurement systems, so it is not necessary to assemble one in the laboratory. These systems are often easy to use, and their performance has been optimized and it is known, so oscillators with really low phase noise can be measured. The downside is that they are not in-expensive. Even when using these specialized systems, it is important to understand the underlying phenomena and the limitations in the system. Unfortunately not all manufacturers describe how their system works. This section will introduce some commercial products, and the used measurement method is given if it has been published.

Wenzel Associates offers three different systems: BP-1000-SC (PLL method), BP-1000-SC (Cross-correlation), BP-1000-RM (Residual) [13]. With these devices carrier frequencies from 5 MHz to 1.5 GHz with offsets up to 100 kHz can be measured with the noise floor down to -190 dBc/Hz.

Agilent Technologies currently offers E5505A Phase Noise Measurement Solution. It is a modular system designed for phase noise measurements, which can be configured to

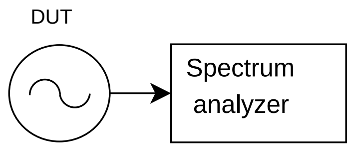


Figure 3.5: Spectrum analyzer method set-up

fulfill the specific needs of a customer. With additional modules the system can measure carrier frequencies from 50 kHz to 110 GHz with offsets from 0.01 Hz to 100 MHz. The noise floor is around -180 dBc/Hz.

3.5.1 Measuring Phase Noise with a Spectrum Analyzer

Phase noise can even be measured directly with a spectrum analyzer. This is not a very good alternative if one wants to measure phase noise close to the carrier frequency, mostly because spectrum analyzers have their own noise properties which can degrade the measurement results. Even in many real life cases the phase noise of DUT is lower than the noise of the spectrum analyzer and thus it becomes impossible to measure the phase noise of the DUT. This method does not require complex outboard measurement set-up, as Figure 3.5 shows. [14, p. 162]

In addition to the limitations in phase noise performance of spectrum analyzers, drifting makes the measurements more problematic. As previously stated, phase noise is measured relative to the carrier and thus it is important to lock the measurement system to the carrier, no matter how much the carrier frequency varies due to noise. In practice the DUT must have a low enough frequency drift when compared to the sweep time of the spectrum analyzer. Otherwise it is impossible to measure phase noise, because the carrier frequency varies too much. In general, synthesized sources have a low enough drift that they can be measured, if the reference oscillator has low enough phase noise. Free running oscillators, on the other hand, usually drift too much to be measured using a spectrum analyzer. [14, p. 162]

Take for example a good 10 GHz cavity oscillator which will drift typically 30 kHz/ min. It takes typically at least 20 s to scan a 50 kHz span with 100 Hz resolution bandwidth. During that time the frequency of the DUT will have drifted 10 kHz. It is thus easy to see that this method has some big limitations. [3] This example was from the 1980s and spectrum analyzers have been improved much since then, but the underlying problem is the same. With the Agilent E4407B spectrum analyzer the same sweep takes 4 s, so the frequency of the DUT would have drifted only 2.5 kHz.

As can be seen from the previous example, with the direct spectrum analyzer measurements care must be taken when deciding the right values for the measurement parameters. In addition, if the resolution bandwidth is too large, the carrier seems to be too wide, its level too high at the envelope detector at offset frequencies, and the measurement re-

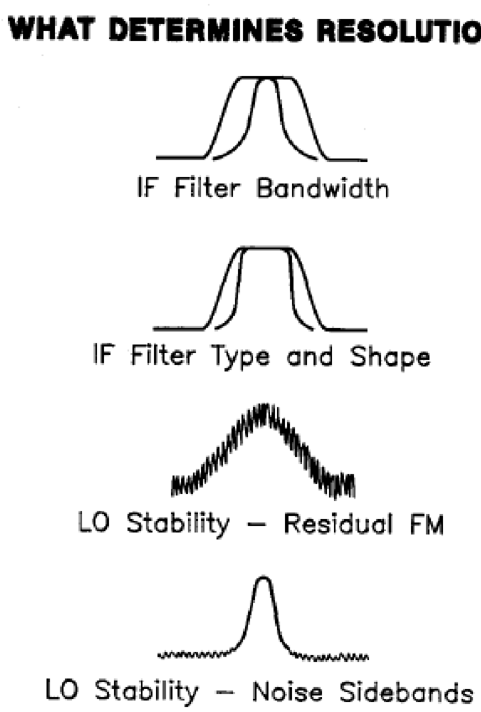


Figure 3.6: What determines the resolution of a spectrum analyzer (Published with permission) [3]

sults are thus misleading. [14, p. 165] Figure 3.6 shows some of those properties which determine the resolution of a spectrum analyzer.

The frequency resolution of a spectrum analyzer depends mostly on the intermediate frequency (IF) filter and the stability of the local oscillator (LO). The resolution bandwidth (RBW) depends thus on the IF filter characteristics. More information about this and other related parameters can be found at [15].

3.5.2 Phase Noise Measurement Modules of Spectrum Analyzers

Some spectrum analyzers are equipped with methods for calculating phase noise during the measurement process. Usually these must be bought separately and in the past they have been expensive. This kind of a phase noise kit is included in the Agilent E4407B spectrum analyzer in the RF-Laboratory. It can measure phase noise directly, or with a spot-frequency type of measurement. In the latter one the analyzer tracks the carrier frequency, and compares its level to the noise level in one offset frequency. This can be used when the carrier drifts too much for the direct measurement, but the biggest problem is that this method is really slow. All interesting offset frequencies must be measured individually and thus the more resolution is wanted to the phase noise plot the slower the measurement process becomes. It takes about 30 seconds to measure one offset frequency. If there are too few measured frequencies, some anomalies in the phase noise plot can be left unnoticed, and with too many frequencies the process takes too long.

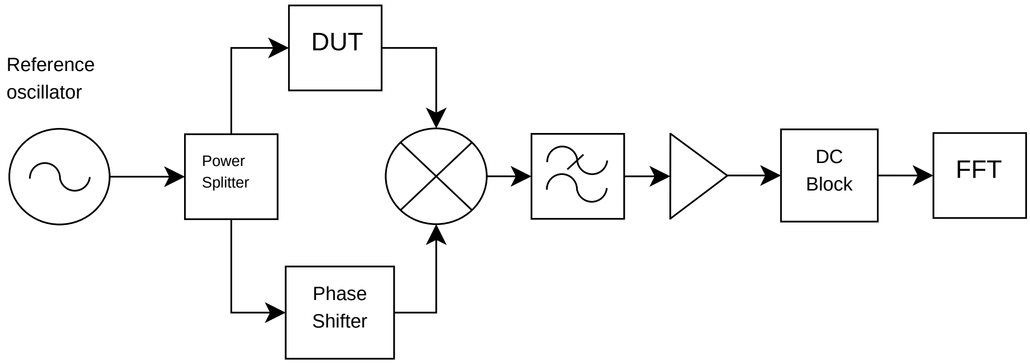


Figure 3.7: Residual method set-up

One alternative for the proprietary phase noise measurement options of spectrum analyzers is a KE5FX GPIB Toolkit written by a radio amateur John Miles, KE5FX. [16] The program runs on Windows, and it can use dozens of different spectrum analyzers controlled through a GPIB bus for making the actual measurements. The more information about this can be found in Chapter 5.

To summarize, measuring phase noise directly with a spectrum analyzer is usually not adequate. It can be used if the performance of the measured oscillator falls in a specific range, but more often than not this is not the case. If the drift of the oscillator is small enough to make the measurement possible, its phase noise is typically so low that it is below the phase noise of the spectrum analyzer and thus the measurements can not be made. Many components in spectrum analyzers add their own phase noise, for example local oscillators, different amplifiers, and mixers. These noise contributions add to each other when performing measurements, and the phase noise performance of a spectrum analyzer includes all of these. On the other hand, if the phase noise of the DUT is so high that measurements could be made, the frequency is usually not stable enough for the measurements to be performed directly with a spectrum analyzer.

3.6 Residual Method

The methods shown thus far can be used to measure only oscillators. There are some methods for measuring two-port devices, and the residual method is one of them. It can be used for example to measure amplifiers, mixers, cables, and filters. Figure 3.7 shows the measurement set-up. In this method the output of a reference source is split with a power splitter. One branch is connected through the DUT to the mixer and the other branch through a phase shifter to the mixer. The phase shifter is adjusted until the phases are in quadrature, and the output of the mixer is measured with a spectrum analyzer. Because the noise from the reference source is coherent at the mixer input and the signals are in the quadrature, it will mostly be canceled out. The remaining phase noise at the mixer output is thus added by the DUT. [17]

4. BUILDING THE MEASUREMENT SET-UP

Time spent planning all the details and parameters applying for the measurement set-up is time well spent. It is important to know the limitations of all the components so that the effect of those limitations can be minimized. It is necessary to check the bandwidth and power specifications because otherwise the signal under investigation may be distorted too much inside the set-up or it may even break something.

This chapter describes the used measurement set-up in more detail. First there is a general description about it, how it works and what kind of parts are required. After that different parts are analyzed more thoroughly and restrictions and problems related to all parts are discussed. After this chapter the reader should be familiar what really is required when building a phase noise measurement system using the PLL method used in this thesis.

4.1 Basics of the Measurement Set-Up

Although the measurement system based on the PLL method is not a really complex one, there are many things which must be considered. The set-up utilizes some commercial off-the-shelf components and the rest were built during this project. Some of the components require a more detailed analysis, and some can be covered with fewer words, but all important parameters are discussed.

As Figure 4.1 shows, the setup consists of two oscillators, a mixer, a low pass filter, and a low noise amplifier. The output of the circuit is measured using a spectrum analyzer. The oscilloscope helps by showing when the system is in phase-lock, but it is not required. Discussion about the different components and their important properties will follow next.

Reference oscillator: The reference oscillator is probably the most important part of the system. With the PLL method it is possible to measure oscillators with equal or worse phase noise performance than the reference oscillator. If the phase noise performance of the reference oscillator is not good enough, its noise can not be separated from the noise of the DUT. In practice, if the difference is more than 20 dB, the noise of the reference oscillator can be neglected. As was discussed in Section 3.2, if the noise levels are close to each other, a correction factor of 0 to 3 dB must be subtracted from the results. This correction factor is described in Eq. (3.1) and Table 3.1. The reference oscillator should also be stable, so that it would not drift too much during the measurements.

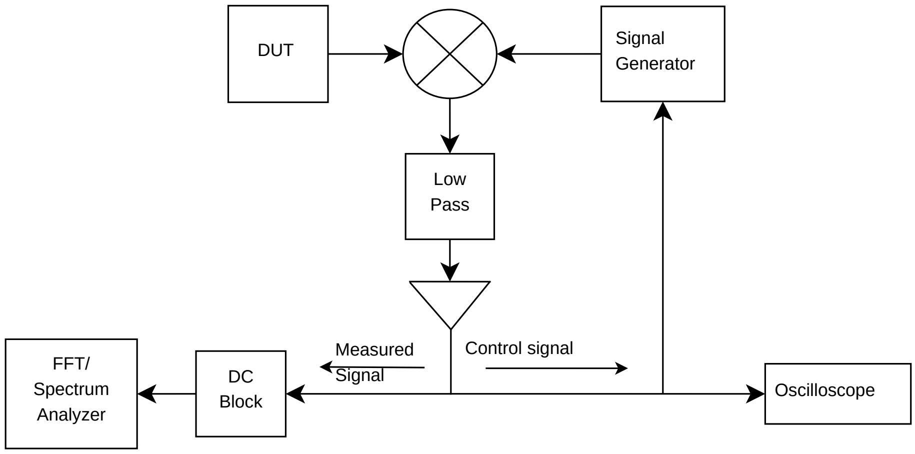


Figure 4.1: PLL method set-up, same as Fig. 3.1, shown here for convenience

Mixer: A basic double balanced diode mixer can be used, so this part is not so critical. The main requirement for the mixer is that it must have a DC coupled port. Double balanced mixers have often better isolation characteristics and the DC offset is also often better than with signal balanced mixers. There are also fewer undesired frequencies at the output of the double balanced mixer than at the output of the signal balanced one. [6]

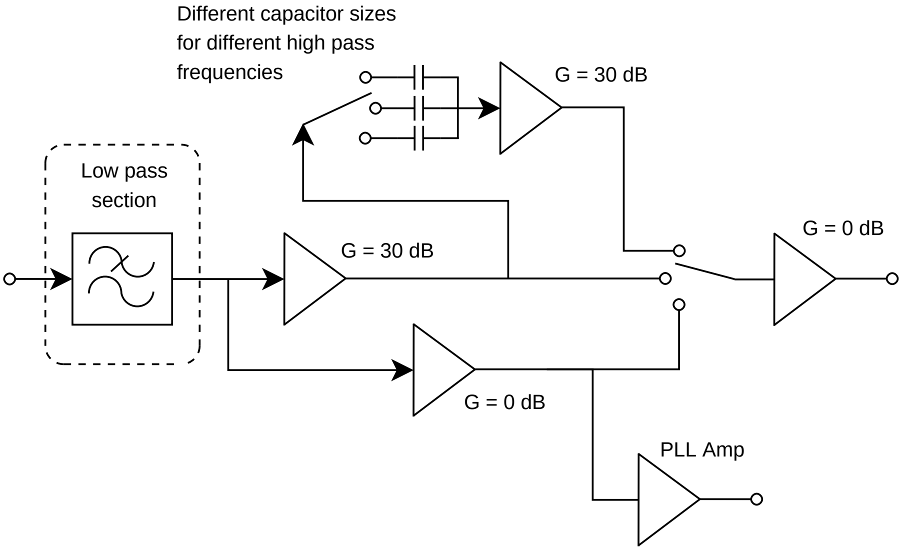
Low pass filter: The low pass filter following the mixer is also not critical, because the frequency of the signal to be measured is so low when compared to the RF frequencies which will be blocked. The performance of the mixer and low pass filter must be known, but it is not necessary to use the latest innovations in their designs.

Low noise amplifier: The low noise amplifier, in turn, is a critical component that must be designed and built well. With careful planning its noise properties are not the limiting part of this measurement set-up. No suitable amplifier was easily available, so one was built during this project. The design and building process is described more thoroughly in the following section.

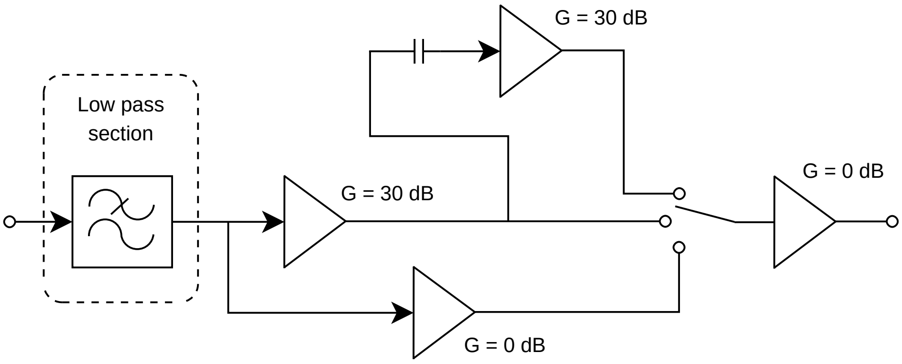
In this project, the Agilent 8648C signal generator was used as the reference oscillator, and Mini-Circuits ZX05-42MH as the mixer. An analog 40 MHz oscilloscope (Kenwood CS-5135) was used as a help in the PLL tuning.

4.2 The Low Noise Amplifier and the PLL

The low noise amplifier (LNA) is used to amplify the weak ripple voltage in the PLL circuit. The bandwidth requirement is from 0 Hz to some hundreds of kHz and it is the most limiting part in this measurement set-up. The bandwidth of the LNA defines what is the highest offset frequency that can be actually measured. In this project the upper -3 dB corner frequency was specified to be between 200 kHz and 300 kHz. The LNA is located right after the mixer and the low pass filter.



(a) Original amplifier



(b) Modified amplifier

Figure 4.2: Amplifier block diagrams

Wenzel Associates Inc. manufactures phase noise measurement systems, and they have provided schematics for a low noise amplifier targeted for this use [10]. The measurement system used in this thesis is somewhat different from the system specified in the document provided by Wenzel Associates, so it was necessary to modify the amplifier circuit to fulfill these needs. Figure 4.2a shows the original block diagram, and Figure 4.2b the modified one.

The block diagrams are almost the same, with only two differences. The modified one has only one high pass option when using the 60 dB option and the PLL amplifier section is also removed. It was decided to drop all unnecessary components, and one high pass option was enough for verifying that this amplifier and whole measurement set-up would work. The PLL amplifier is not required in the set-up used in this thesis, because the signal generator can be tuned with the signal from the main output of the LNA. If the signal generator had been replaced by a voltage controlled oscillator (VCO), the PLL amplifier would have been necessary. In that case the output of the LNA is not adequate, because usually the tuning voltage is 0 to 5 V or something similar. In the original schematics the

Table 4.1: Prices of the components in the LNA

Component	Amount	Total price / €
SK170 FET	2	1
2N5639 FET	1	0.1
Resistors, metal-film, SMD	9	0.9
Capacitors, SMD	8	0.4
Capacitors, electrolytic	2	0.5
Inductor	1	1
AD825 operational amplifier	1	3
LM833 operational amplifier	1	2
DPDT Switch	2	6
Connector, BNC	2	3
Connector, SMA	1	1.5
PCB, FR4	5 cm x 8 cm	1
Total price		19.4

output of the PLL amplifier was between the used supply voltages ($\pm$15 V), so it would have been necessary to alter the amplifier circuit also in this case. The amplifier used in this thesis could also have been converted to output different tuning voltages, but due to time limits it was not done.

Component level modifications were also necessary because not all components were easily available. For example, the SK170 FETs was replaced by the 2SK369 FETs and the AD825 operational amplifier was replaced by the 310 operational amplifiers. The final schematics and layout can be found in Appendix A. Table 4.1 lists the approximate prices of the used components.

Basically the chosen amplifier circuit is a high-gain low noise amplifier with three different gain and bandwidth options. The low pass section was modified and first created as an external version to make it easier to measure the performance of the system. In the end the low pass section was included in the final version of the PCB board.

It is not very difficult to build a suitable amplifier with low noise properties, even when using ordinary operational amplifiers, although care must be taken when choosing the different parts. For example carbon film resistors are not suitable because they have poor noise properties, but metal film and wirewound resistors are good candidates. Operational amplifiers and FETs have really varying noise properties, depending solely on their intended use, and this must also be considered when choosing the parts. [10]

Basically this amplifier is a basic voltage amplifier, but it has some features which make it good for this use. For example the amplifier has three different gain settings which can be used to avoid overloading the amplifier and blocks following it. Another way to avoid overloading devices after the amplifier is to use switchable attenuators, but it is good to have alternatives. Sometimes one method will work better than the other, as was noticed when performing the actual measurements.

Switchable attenuator or switchable gain is necessary because of the huge dynamic range of the measured signal. As already stated, the required dynamic range can exceed 120 dB, and usually the usable dynamic range of a spectrum analyzer is somewhere around 70 to 90 dB. In this context dynamic range is defined as: "The ratio, expressed in dB, of the largest to the smallest signals simultaneously present at the input of the spectrum analyzer that allows measurement of the smaller signal to a given degree of uncertainty." [15] It is thus necessary either to attenuate the carrier or amplify the noise. The actual measurement procedure is described in more detail in section 5.4.

The amplifier has three different bandwidth/gain-related settings. The 0 dB and 30 dB gain settings have no additional high pass filter, so it is possible to measure small offset frequencies. The 60 dB gain setting has an additional high pass filter, a 1 μF capacitor in series, so that low phase noise far from the carrier can be measured even when strong noise is present near the carrier. With the 1 μF capacitor the -3 dB frequency should be around 250 Hz [10]. Here the -3 dB frequency means the frequency where the amplifier gain has dropped 3 dB from the highest value in the frequency response figure. The actual performance of the real amplifier is verified in Section 5.2.

In this thesis, the EXT DC modulation mode of the Agilent 8648C is used for the PLL as described in [5, p. 9.11]. When the signal from the mixer is connected to the MOD INPUT/OUTPUT connector and the EXT DC modulation is turned on, the frequency of the output signal of the signal generator is varied relative to the DC voltage at the connector. Thus the signal generator can work as a part of the PLL circuit, because the output voltage of the mixer is relative to the phase difference of the two oscillators. When the signals at the mixer inputs are in quadrature phase, the output is 0 V DC. This voltage varies to either to positive or negative direction depending solely on the phase difference between the input signals. The DC voltage is used to tune the frequency of the reference oscillator to get 0 V at the input of the signal generator.

Figure 4.3 shows the final amplifier construction. SMA was used as the input connector, and BNCs for the outputs. This amplifier is not an RF amplifier, so there is no real need to use SMA connectors, but the mixer output connector is a SMA and thus no adapter or special cable is required.

There are two switches in the LNA for the gain changes. The upper one is for choosing between a buffer and amplification. When the upper switch is down, the amplifier acts as a buffer, and when it is up, the amplifier amplifies the signal. The lower switch chooses the gain setting. At the right position there is 30 dB of gain and at the left position there is 60 dB of gain. Figure 4.4 shows the different switch positions and how they control the gain settings.

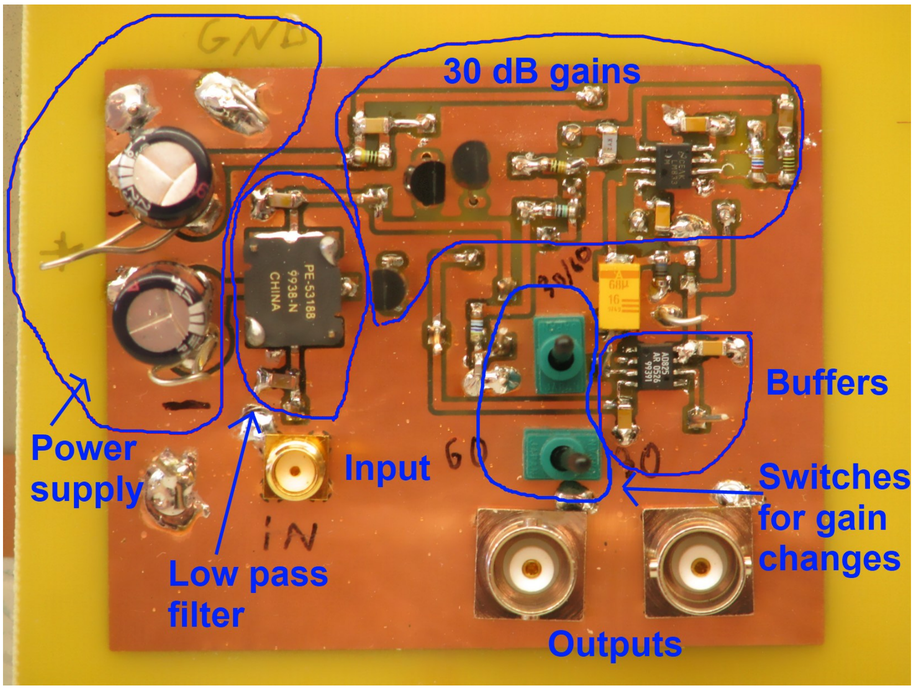


Figure 4.3: The PCB of the low noise amplifier

(a) Buffer (b) 30 dB gain (c) 60 dB gain

Figure 4.4: Switch settings

4.3 Spectrum Analyzer

Care must be taken when deciding which spectrum analyzer to use in this measurement set-up. Not all RF spectrum analyzers can be used. The biggest problem is that the frequency range of the spectrum analyzer must be suitable and it may prohibit using the particular analyzer. Take for example the bandwidths of the two spectrum analyzer which can be found in the TUT RF Laboratory: Hewlett-Packard 8595E 9 kHz. . . 6.5 GHz and Agilent E4407B 100 Hz. . . 26.5 GHz. E4407B has two coupling options, 'AC' with an internal DC block, and 'DC' with no DC blocks. With the AC coupling the frequency range starts at 10 MHz and with the DC coupling it is crucial to remove any DC voltage before the signal gets to the spectrum analyzer because it does not withstand any DC voltage in its RF input.

The E4407B spectrum analyzer may be usable if there is nothing else available and the restrictions have been taken care of, but it would be better to find a spectrum analyzer designed for audio frequencies. One additional big problem is that the level at the input of the spectrum analyzer can become too high if the PLL becomes unlocked and the beat note appears. The level difference between the carrier frequency power and noise power can exceed 120 dB even with many VCOs and usually the usable dynamic range of a spectrum analyzer is much less than that. Attenuators must thus be used when measuring the level of the carrier and they must be taken off when measuring the noise level. The carrier level is measured from the beat note, and this is described more clearly in section 5.4.

The amplifier used in this thesis can deliver a signal with approximate amplitude of 9 V_{RMS} to its output when the 60 dB gain setting is used, and the input of E4407B can tolerate only $+30$ dBm = 1 W. Even though the output impedance of the amplifier is not 50 Ω, the input impedance of the spectrum analyzer is and the performance of the spectrum analyzer will be the limiting part in this system, so 50 Ω is thus used here. The output impedance of the amplifier is low, so it should be able to deliver its full output voltage to the 50 Ω load. The maximum output level of the LNA for the 50 Ω load can be calculated as

$$P_{\mathrm{AmpOut}} = \frac{U^2}{Z_0} = \frac{(9\ \mathrm{V_{RMS}})^2}{50\ \Omega} \approx 1.62\ \mathrm{W} \approx 32.1\ \mathrm{dBm} \tag{4.1}$$

Agilent does not state the input level which will break the spectrum analyzer, and probably 32.1 dBm would not be absolutely harmful, but it is still better not to exceed the stated maximum input level $+30$ dBm. The maximum input level is $+30$ dBm when all the internal attenuators are set to their maximum settings. Another thing to consider is that the level at the mixer inside the spectrum analyzer should be under -10 dBm. If higher input levels are present, input attenuators inside the spectrum analyzer should be used. That is one reason why the amplifier used in this system was built with switchable gain settings. When using the 30 dB gain the output should be under the limits, so with that setting E4407B could be used without any problems. Still some caution is required during the measurement process.

When measuring phase noise, the most interesting frequencies are usually between 0 Hz and 100 kHz relative to the carrier frequency, and in the PLL method the carrier is mixed to 0 Hz, so it is thus the lowest interesting frequency. It is not crucial to have the bandwidth starting exactly from 0 Hz, but it can be high pass filtered a little higher. In this set-up the -3 dB corner frequency is around 20 Hz. The highest interesting frequency, on the other hand, can be little higher in some cases. In E4407B datasheet, phase noise is shown even up to 10 MHz. The upper -3 dB corner frequency of the amplifier used in this thesis is about 250 kHz, so that is the practical limit in this case. A more limiting part may be the spectrum analyzer used. If an audio spectrum analyzer is used, the upper limit of the bandwidth is probably around 100 kHz. If an RF spectrum analyzer is used, it will not limit the upper frequency limit.

4.3.1 About different Spectrum Analyzers

During this project four different spectrum analyzers were tested, and it was found that the E4407B was the most suitable one, despite the previously stated shortcomings. The Advantest R9211A was tested first, because it is designed for audio measurements and its bandwidth (0 to 100 kHz) was suitable. There were some problems with this particular instrument which made it not suitable for this use. Perhaps the biggest problem was that it was not possible to get the measured data to a computer for processing. The measured trace could be printed or saved only in a binary format to a 720k diskette. Because it was a binary format and there was no program available for reading it, it did not help at all. Another problem was that this device was really slow for making any measurements, because it was old and lacked processing power.

Also a basic laptop computer with two different external sound cards was tested. One of the cards had a 48 kHz sampling frequency so it could reach a little over 20 kHz in the measurements, which is far from enough in this set-up. This card was used in the beginning to verify that the built PLL method set-up works at all. Another sound card had a 96 kHz sampling frequency, so it could reach to over 45 kHz which is still low, but can be used if nothing else is available. There are also sound cards with a 192 kHz sampling frequency, which results in adequate bandwidth of about 95 kHz. That kind of a card was not available, so it was not tested during this project. The measurements with the USB sound cards were made using the Smaart v.6 as the analyzer [18]. It is designed for acoustical measurements, but can be used for any raw spectrum measurements up to the highest frequency available from the analog to digital converter.

4.3.2 Problems in using a Laptop Computer as a Spectrum Analyzer

There are some big problems when using a laptop computer based measurement system, but they can be solved if the used devices and measurement circuits are designed properly. Laptops in general exhibit really poor performance in audio measurements. Their power supplies and internal grounding schemes are not designed for interconnecting to complex systems. A complex system can be as simple as a laptop and a signal generator separated by some meters and connected to different power outlets. Usually the internal sound cards are also not suitable for measurement work.

This noisy sound card problem can usually be disregarded when using external sound cards which connect to a computer via USB or Firewire. It sometimes is also necessary to disconnect the power supply of the laptop when making the actual measurement because when doing that the noise floor of the computer may be lower, and nasty ground loops may be avoided. This power supply problem was seen once during the project, but it could not be repeated, so no data is available. Probably the used two laptops were not so problematic ones.

5. TESTING THE SYSTEM

A system without any performance verification is not very useful. If the performance of all components is not verified, it is impossible to know whether the measurement results can be trusted or not. The amount of details which must be gathered and verified from a specific component varies a lot, but the engineer must know what is required.

In this chapter the components used in the measurement set-up are measured. Their performance is verified and compared to known standards or published performance specifications if possible. First there is some discussion about measurement basics, followed by a description of all the preparative measurements where the performance of the different parts was verified, and finally there are the results from the actual phase noise measurements. For each block separately, it is described which parameters are essential to measure and why, and what problems they can cause to the final results. At the end of this section can be found some discussion about the uncertainties in the PLL method.

5.1 Measurement Basics

Phase noise is measured relative to the carrier, as already stated, and thus the difference between the levels is important, not the absolute signal levels. Signal levels must still be roughly known, because there are limitations in the measurement equipment. It must be verified that the signal does not exceed the maximum input level of devices. It should also be verified that the signal level is not too low, because thermal and instrument related noise is usually a bigger problem with low signal levels than with high.

First there will be some preparative measurements where all the used equipment is checked and performance verified. Some parts of the equipment were calibrated by an accredited calibration laboratory in August 2009, so their performance was taken as granted. After those performance checks, the real measurements will follow.

Table 5.1 shows the oscillators which were measured in this thesis. From now on, their respective characters will be used for referencing to a particular oscillator. The oscillators A–C were made as student projects at the Department of Electronics. Oscillators D and E are Agilent signal generators. Oscillator E offers two different oscillator modes. In Mode 1 the phase noise performance is optimized for offset frequencies under 10 kHz and in Mode 2 it is optimized for offset frequencies over 10 kHz. These two modes will also be used in different measurements, and it will be stated which one was used in each case. The modes will be called E1 and E2, respectively. Figure 5.1 shows the published phase noise

Table 5.1: Oscillators measured in this thesis

#	Frequency (MHz)	P_{out} (dBm)	Designer / Model
A	1800	-3	Student project 1
B	422	-15	Student project 2
C	420	8	Student project 3
D	0.009-3200	$-10/-5$	Agilent 8648C
E	0.25- 3000	$-10/-5$	Agilent ESG-D3000A

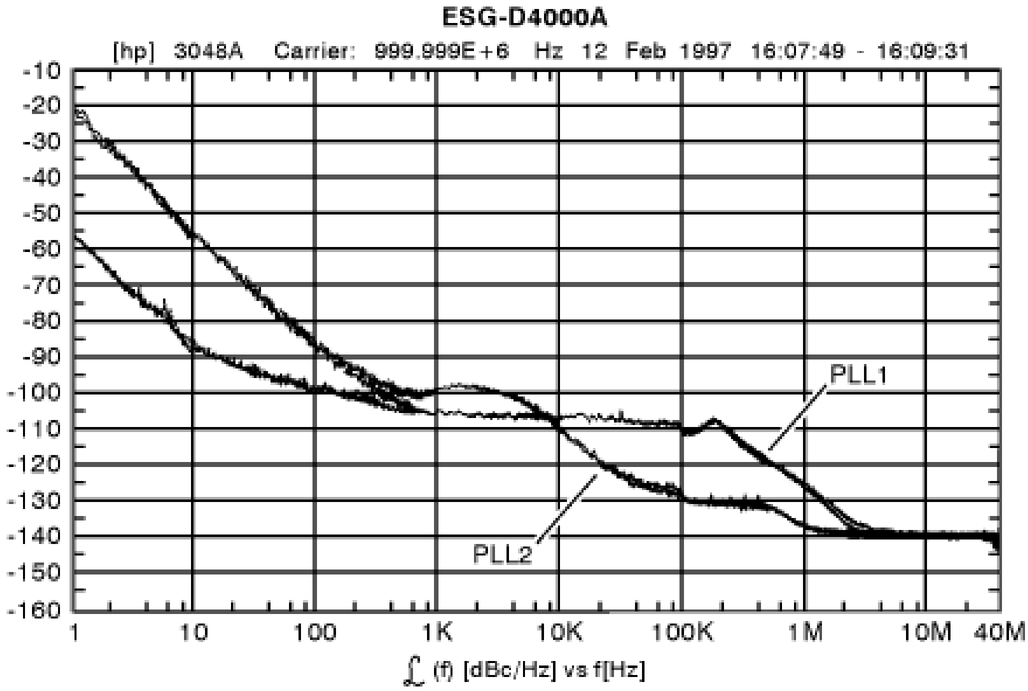


Figure 5.1: The published phase noise performance of the ESG-D4000A (Published with permission) [19]

specification of ESG-D4000A. Figure 5.2, in turn, shows the phase noise curves measured with the PLL method during this project. Even though the published specification is from a different model, the phase noise properties are probably similar within the same series. In addition, the published specification only tells the typical performance of this series, and there can be some differences in the actual measured performance. The actual phase noise performance is probably better, because it is unwise to publish a specification which is much better than the actual performance of a device.

5.2 Verification of the Measurement Set-up

When verifying that the measurement setup works as designed, it is necessary to measure all components individually and then gradually connect them together and verify that all the components work together as a system. This section describes the verification process. It is divided in parts based on the amount of measurements and details related

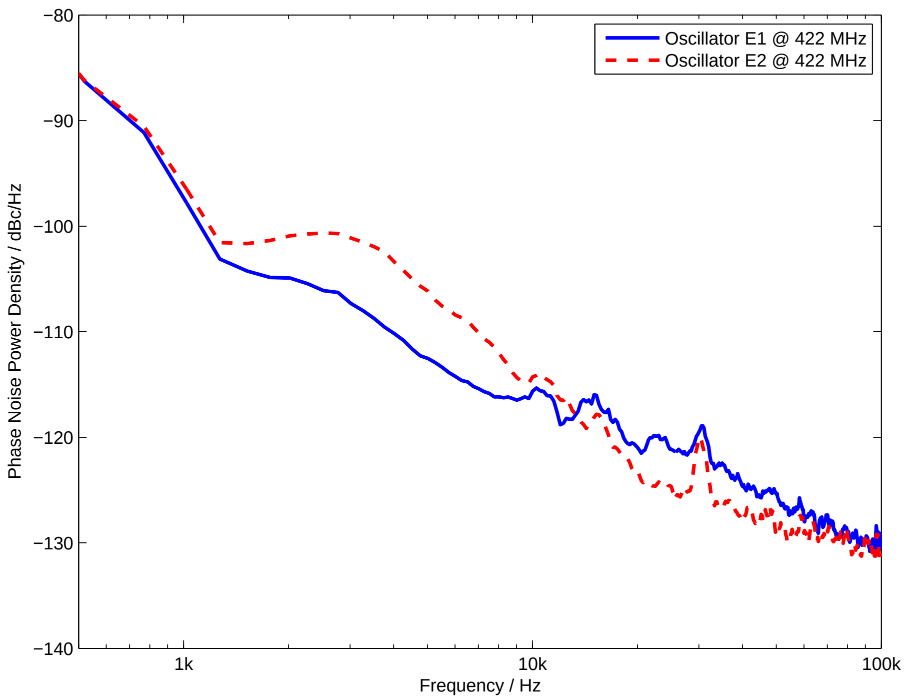


Figure 5.2: Measured phase noise when using different PLL modes of the ESG-D3000A

to the measured components. After the verification process described in this section, the whole set-up can be built and final measurements made.

5.2.1 Agilent E4407B Spectrum Analyzer

First measurements were made with a Agilent E4407B spectrum analyzer and two signal generators D & E using the measurement set-up shown in Figure 3.5. The phase noise module of the spectrum analyzer was compared to the phase noise module of the KE5FX GPIB Toolkit [16], which was already described in section 3.5. The results seemed to be too close to each other, with only the 3 GHz trace a little different, as shown in Figure 5.3. 3 GHz is the highest available frequency of ESG-D3000A, so it is not surprising that the performance is poorer than in the middle of the frequency range. Both the signal generators have too good phase noise properties, to be measured with the spectrum analyzer, causing the phase and other noise properties of the spectrum analyzer to dominate. The conclusion is that this method can not be used to measure these oscillators. The phase noise figure of E4407B can be seen in Figure 5.4 and when comparing that figure to the measurement results in Figure 5.3, it is easy to see that the spectrum analyzer is the dominant noise contributor. Later, more noisy oscillators were measured directly with the spectrum analyzer, and the results from these measurements can be found in section 5.3.

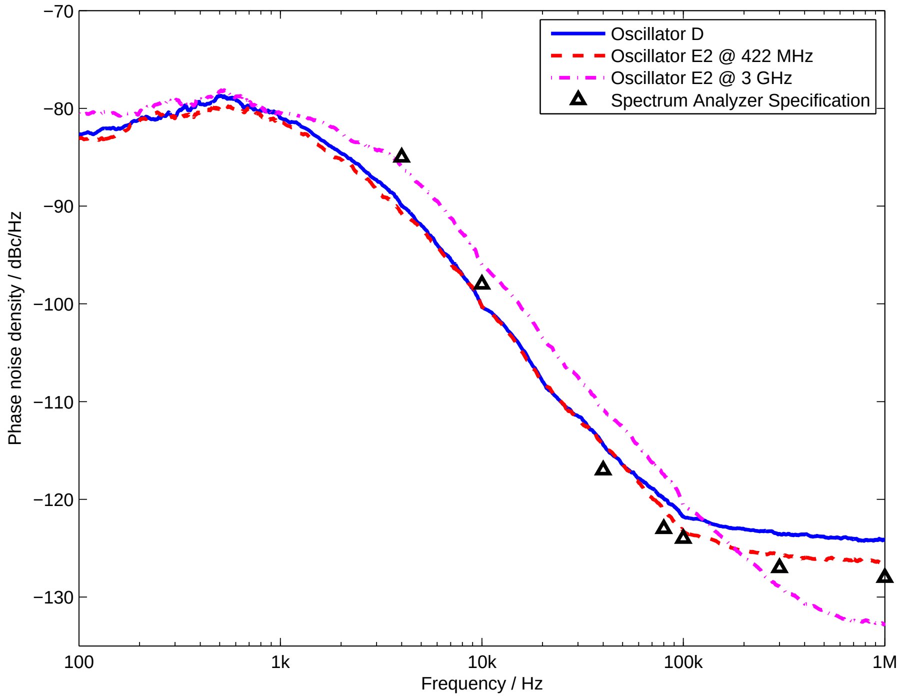


Figure 5.3: Phase noise of different oscillators, measured with the phase noise module of E4407B. Δ-points show the published phase noise specification of E4407B [20].

5.2.2 Mini-Circuits ZX05-42MH Mixer

Figure 5.5 shows the measurement set-up for verifying that the mixer would be suitable. Agilent 8648C was used as the oscillator. Its output was fed to the power splitter (Mini-Circuits ZN2PD-9G) and then through two cables with different lengths to the mixer. These cables caused different amounts of delay to the signals. There would thus be some phase difference between the two signals at the mixer inputs resulting in a DC voltage at the output. The DC voltage was measured with a basic voltmeter. A real phase shifter would have been of help in this set-up, but it was not available. With such a device the phase could have been adjusted more gently, but the cable method worked as well for this basic check. At this point the only interesting part is whether the output would be a DC signal relative to the phase difference of the input signals or not.

The electrical lengths of the cables were measured with the Agilent 8722D vector network analyzer and they were 2.7 ns and 8.8 ns. The DC level was 89.43 mV when the frequency was 1809 MHz. The output power of the signal generator was 0 dBm during these measurements. 1809 MHz was chosen for the test frequency, because the bandwidth of the power splitter is 1700 to 9000 MHz and 1809 MHz offered the biggest DC level near the lower limit of the power splitter.

It is important to have a rough idea of the power levels in different parts of the measurement setup. With 0 dBm = 1 mW output power and 50 Ω system the output voltage

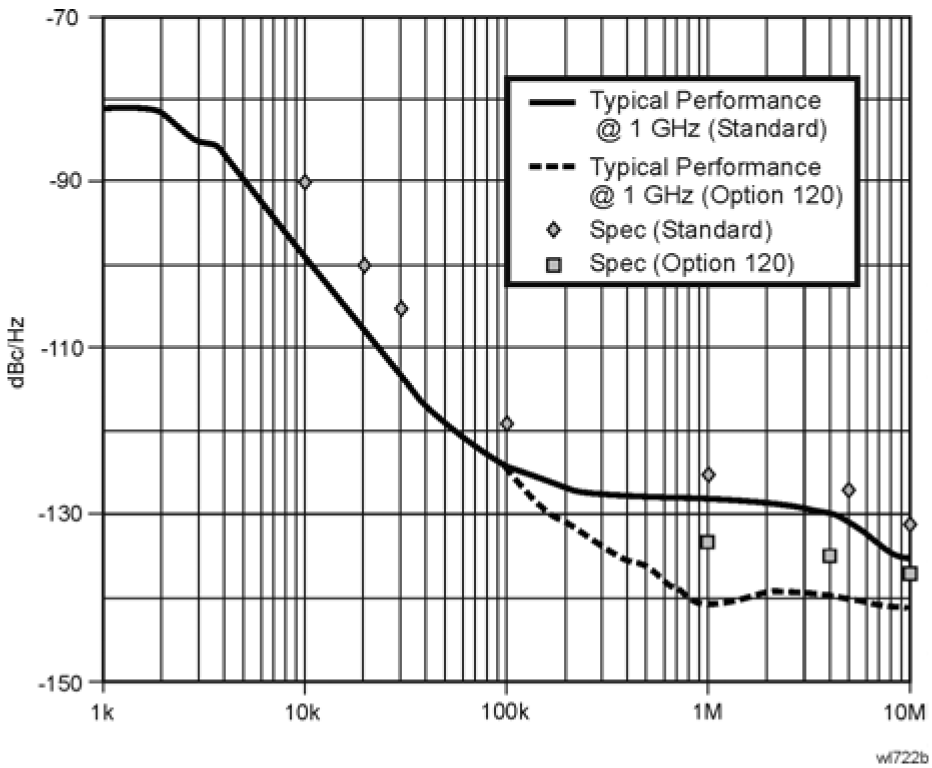


Figure 5.4: Phase noise specifications of Agilent E4407B (Published with permission) [20]

of the signal generator is:

$$U_{\text{RMS}} = \sqrt{PZ_0} = \sqrt{(1\ \text{mW})\,(50\ \Omega)} \approx 224\ \text{mV} \tag{5.1}$$

Because the signal goes through the power splitter it is attenuated to -3.4 dBm ≈ 0.457 mW [21] and thus at the input of the mixer the signal levels are (the cable attenuation is approximated to be zero):

$$U_{\text{RMS}} = \sqrt{PZ_0} = \sqrt{(0.457\ \text{mW})\,(50\ \Omega)} \approx 151\ \text{mV} \tag{5.2}$$

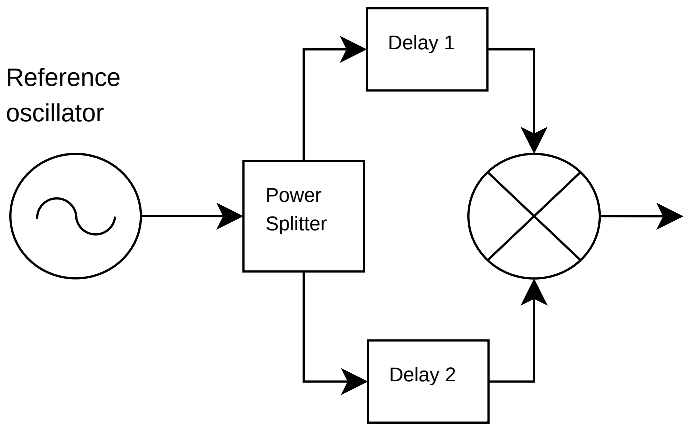


Figure 5.5: The measurement set-up used to measure the mixer DC level

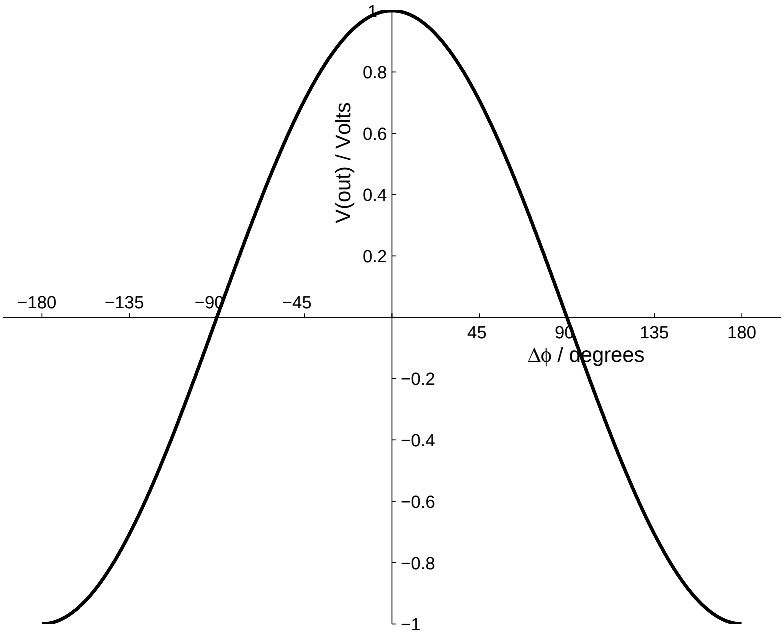


Figure 5.6: Normalized mixer output voltage vs. phase difference of inputs

Because of the losses, mostly the conversion loss of the mixer, the DC level at the output of the mixer is lower than the level of the input signals. The conversion loss can be obtained from the specification sheet of the mixer, but it is not suitable for these calculations, because of the way it is defined: The specification assumes that the level of the LO signal is much higher than the level of the RF signal and that there is a difference in the frequencies. In the measurement set-up used in this thesis, the levels should be similar and the frequencies are the same.

If the output level of the mixer to the 50 Ω load is $V_{\mathrm{RMS}} = 90$ mV as was measured with the voltmeter (high impedance input), the output power of the mixer is

$$P_{\mathrm{out,mixer,dBm}} = 10\log_{10}\left(\frac{\frac{U^2}{Z_0}}{1\ \mathrm{mW}}\right) = 10\log_{10}\left[\frac{(90\ \mathrm{mV})^2}{(50\ \Omega)\,(1\ \mathrm{mW})}\right] \approx -7.9\ \mathrm{dBm} \quad (5.3)$$

The power level has dropped about 4.5 dB inside the mixer.

Figure 5.6 shows the theoretical output of an ideal mixer with varying phase difference between the input signals. The figure is normalized, so that the maximum output level is 1 [22]. The exact DC level is not important in this measurement method, because the DC is filtered away before the measurements and the interesting part is the ripple in the DC voltage, which is related to the phase noise of the oscillator. The ripple was measured with an oscilloscope, and it was from some millivolts to a couple of dozens of millivolts, so it is a low level signal. The ripple was observed using the same oscillator as a source and varying its output frequency, so the behavior of the ripple could be observed with the varying phase difference between the input signals.

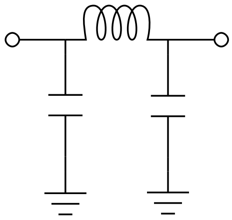


Figure 5.7: The schematics of the Butterworth low pass filter

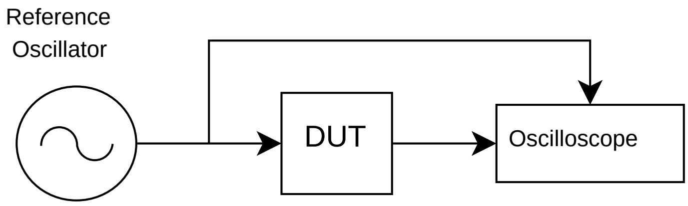


Figure 5.8: Measurement set-up used for measuring the (amplitude) response of the low pass filter, the attenuators and the LNA

5.2.3 Low Pass Filter, Attenuators, Cabling and Adapters

The high frequency content of the signal after the mixer must be removed. The interesting part of the signal is between 0 Hz and about 1 MHz. With the carrier frequencies in the gigahertz range, it is not necessary to have a sharp cutoff right after the highest frequency of interest. It is thus a good idea to build a filter with only a few components and thus with few non-idealities. The pass band of the filter must also be flat.

With these specifications in mind, it was decided that a Butterworth LC-type low pass filter would be the right one. Figure 5.7 shows the schematics of the chosen filter. The filter was built around a 2.2 µH inductor which was available and had a suitable form factor for mounting on the PCB. Some calculations for the required capacitor values were made with a program offered by Dale Heatherington [23]. The calculations showed that with the available inductor and 454 pF capacitors, the -3 dB corner frequency should be around 7 MHz. The frequency is a little bit high, but it was still decided to build the filter with these components.

The frequency response of the filter was measured with the set-up shown in Figure 5.8. The test signal is a sine wave of the required frequency. The first channel of the oscilloscope measures the signal level before the DUT and the second one after the DUT. The signal frequency is varied inside the chosen frequency range and both the input and output levels of the DUT are measured. From the results the gain is calculated and converted to decibels. This same set-up was used for measuring the attenuator settings and the frequency response of the LNA. Because the lowest available frequency of the RF signal generators in the RF Laboratory was still too high, the Thurlby-Thandar TG230 function generator with a more suitable frequency range was used.

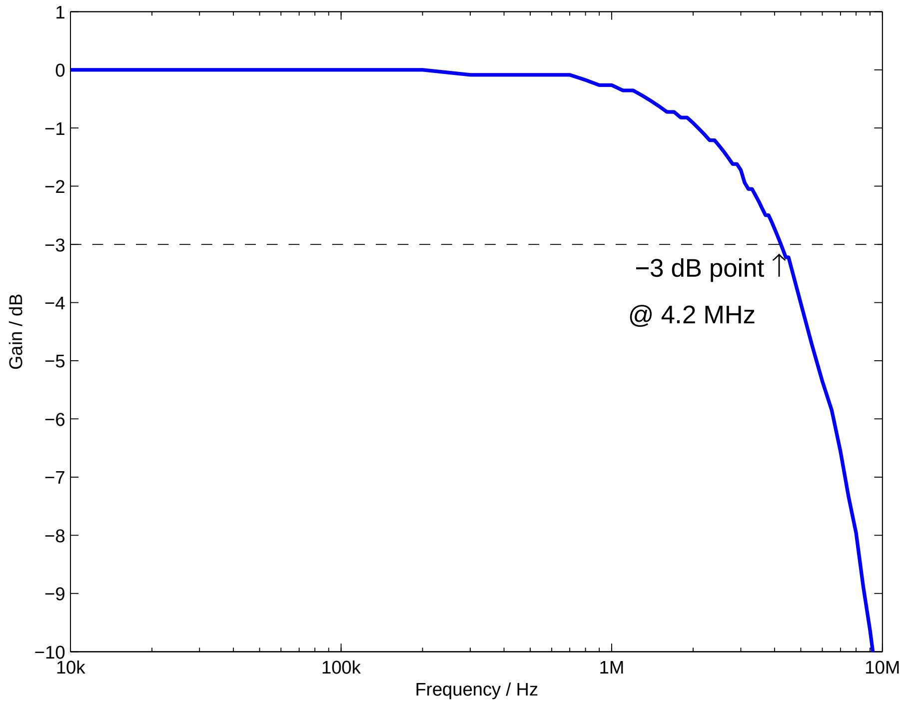


Figure 5.9: Measured frequency response of the low pass filter, −3 dB point at 4.2 MHz

Table 5.2: Measured performance of the attenuators used

#	Atten. / dB	Bandwidth / GHz	−10 dB	−20 dB	−30 dB	−40 dB
A	0–110	0–18	−9.9	-	−29.3	−36.7
B	0–110	0–4	−10.8	−20.1	-	-

According to measurements, the filter was suitable for use in the measurement set-up. As Figure 5.9 shows, the −3 dB corner frequency is lower than was designed, around 4.2 MHz, but it is actually better this way. There are various reasons for this drop in the corner frequency. The values of the different components are not exact, but these variations alone can not alter the frequency as much as it changed. The filter was designed for 50 Ω source and load impedances, and this is not the situation with the real system. The source impedance is around 50 Ω because the source is the mixer, but the load impedance is something totally different. The exact value is not known, but is quite high because of the voltage amplifier input that follows the low pass filter. In addition, the performance of the filter was measured with an oscilloscope with an input impedance around 1 MΩ + X pF.

Some attenuators were also used in the different parts of the measurement system and their settings verified. Table 5.2 shows the attenuators used and their measured specifications at different attenuator settings. The measurements were made with the same set-up as shown in Figure 5.8.

There are no specific requirements for the cabling or adapters. All cables and adapters from the RF Laboratory can be used. Because the set-up utilizes devices and components with different connectors, it is important to plan the cabling in a way that minimizes the number of adapters. The locations of various devices and the routing of cables should also be considered to minimize errors caused by external electro-magnetic fields and cross-talk.

5.2.4 Low Noise Amplifier

Figure 3.1 shows that the PLL method includes an LNA. It is located after the mixer and the low pass filter. The amplifier was described in more detail in Section 4.2 and in this section the amplitude response of the amplifier is verified.

The amplitude response of the amplifier was measured with two different methods. The first method is shown in Figure 5.8. The second method uses broadband noise as the test signal and the Agilent 4407B spectrum analyzer as the measurement instrument. The noise level is measured with the gain of the amplifier set to the different settings (0, 30 and 60 dB). The actual gain and the frequency response can be obtained by calculating the difference in level between 0 and 30/60 dB gain settings. This method assumes that the 0 dB setting has no gain or attenuation. The response was measured with the external high-pass filter, because the high-pass filter affects the frequency response of the amplifier circuit.

Figure 5.10 shows the results from the amplitude response measurements. The black dashed lines show the -3 dB thresholds. There are two traces for both gains, because the frequency response was measured with two different methods. Due to measurement restrictions in the second method, caused by the Agilent E4407B spectrum analyzer, results from this measurement start at a somewhat higher frequency.

The 0 dB setting was not measured very thoroughly here, because it is only a buffer, and its frequency response is assumed to exceed the other limiting parts. Quick verifications confirm this assumption: it works as a 0 dB buffer at least up to 1 MHz.

Figure 5.11 then shows more clearly how flat the response is in the bandwidth of 0.5–100 kHz. This range was chosen for closer investigations because it is the main range for the phase noise measurements. It is the range where the amplifier has the flattest response and thus there is no need to make any corrections for the measured phase noise because of non-flat gain of the amplifier. The trace shows the difference to the mean value of the gain in the chosen region, at specific frequencies. It can be seen that the gain starts to decrease when approaching 100 kHz, but because the ripple stays under ± 0.5 dB, it can be neglected.

As Figure 5.10 shows, the frequency response of the amplifier is close to the specifications. The high-pass -3 dB frequency was specified to be around 20 Hz and this is verified in the 30 dB trace. Also the lower -3 dB frequency of the 60 dB is close to the specified one, which was 250 Hz. In those frequencies which are covered by both the oscilloscope and the spectrum analyzer measurement results, the traces are so close to

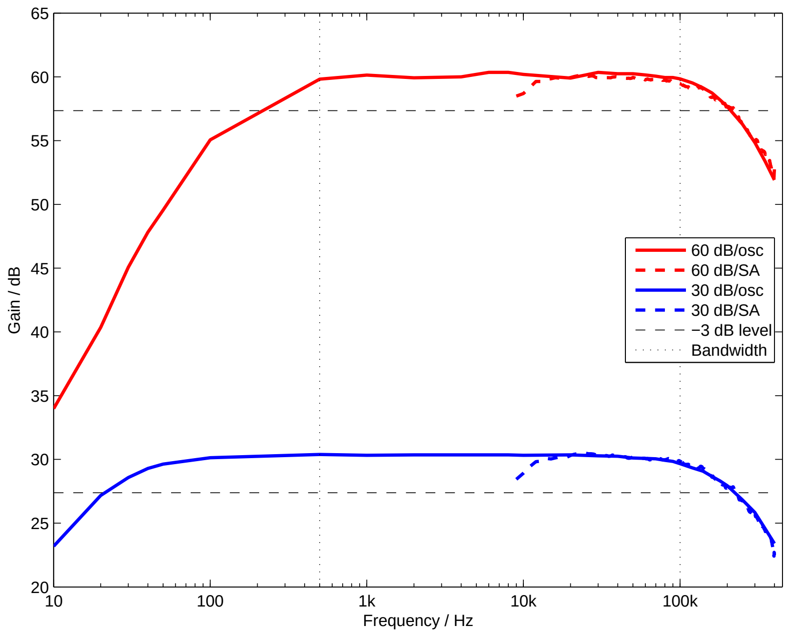


Figure 5.10: The frequency response of the amplifier. Osc means that it was measured with the oscilloscope method and SA that it was measured with the spectrum analyzer method. Bandwidth means the bandwidth which was chosen for closer investigations shown in Figure 5.11.

each other that the results seem reliable. In the 30 dB trace the biggest difference is about 0.3 dB and in the 60 dB trace it is about 0.4 dB.

5.3 Phase Noise Measurements with the Spectrum Analyzer

One of the free running oscillators (B) was stable enough to be measured with the phase noise measurement utility of E4470B and KE5FX GPIB Toolkit [16]. It was measured three times with the cabling and other settings staying the same all the time. Figures 5.12 and 5.13 show the results. As the figures clearly show, it is hard to do the measurements accurately when the measured oscillator is a free running one. There are big variations in the 0.1–10 kHz region and from 5 kHz up the performance of the spectrum analyzer degrades the accuracy, as can be observed from Figure 5.4. The variations in low frequencies depend probably on the stability of the carrier frequency during the measurement period. Sometimes it varies more and sometimes less, and because it is a random process by nature, no two traces will be similar.

As was already seen in Figure 5.3, the phase noise performance of the spectrum analyzer is not good enough for serious measurements. This spectrum analyzer method is an easy method to use when the DUT falls in the right performance range so that it can be measured. At least with the E4407B spectrum analyzer used in this project the range just is so small that almost no oscillator falls in it. Either the phase noise performance of the oscillator is too good, or the frequency of the oscillator drifts too much. In both cases the result is that phase noise cannot be measured with this method.

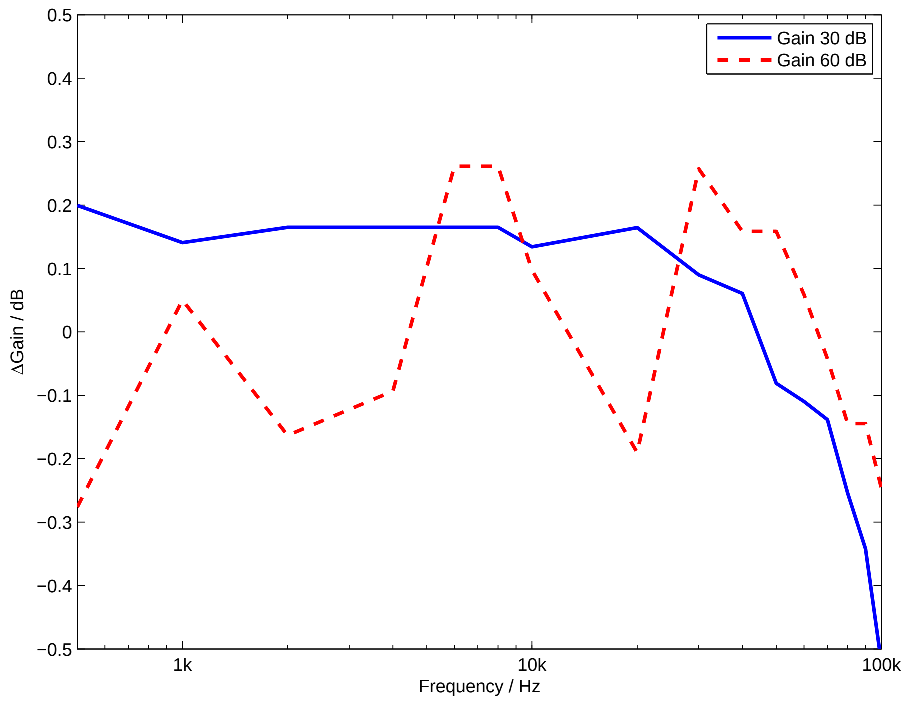


Figure 5.11: The variation in gain compared to the mean gain in the shown bandwidth

The performance of some oscillators was measured also with the spot-frequency mode of the phase noise module of the spectrum analyzer. The results are shown in Figure 5.14. Phase noise was at first measured only up to 100 kHz, because that is the upper limit of the PLL method. There was an anomaly at 100 kHz so it was decided to extend the range to 150 kHz to observe the behavior of the method in larger offset frequencies. The spot frequency mode works a little better than the direct phase noise measurement method of the E4407B when the DUT is more noisy, even though there is a defect with the measurement module. The module changes resolution bandwidth (RBW) at specific offset frequencies, and the measurement module does not take this into account. It assumes that the RBW stays the same all the time. The RBW changes at 30 kHz and 100 kHz offset frequencies, and at larger offset frequencies a correction factor must be subtracted from the measurement results. It can be calculated from the following equation [24]

$$H = 10\log_{10}\left(\frac{\text{RBW}}{1\ \text{Hz}}\right) \tag{5.4}$$

The H values with different RBW values must be calculated, and the the smallest one subtracted from the other values. This is the correction factor which must be subtracted from the results in larger offset frequencies. So, the RBWs are 1 kHz, 3 kHz, 10 kHz for offset frequencies up to 30 kHz, 100 kHz, and 150 kHz respectively. H values are thus 30 dB, 34.77 dB, and 40 dB. With offset frequencies up to 30 kHz nothing is subtracted.

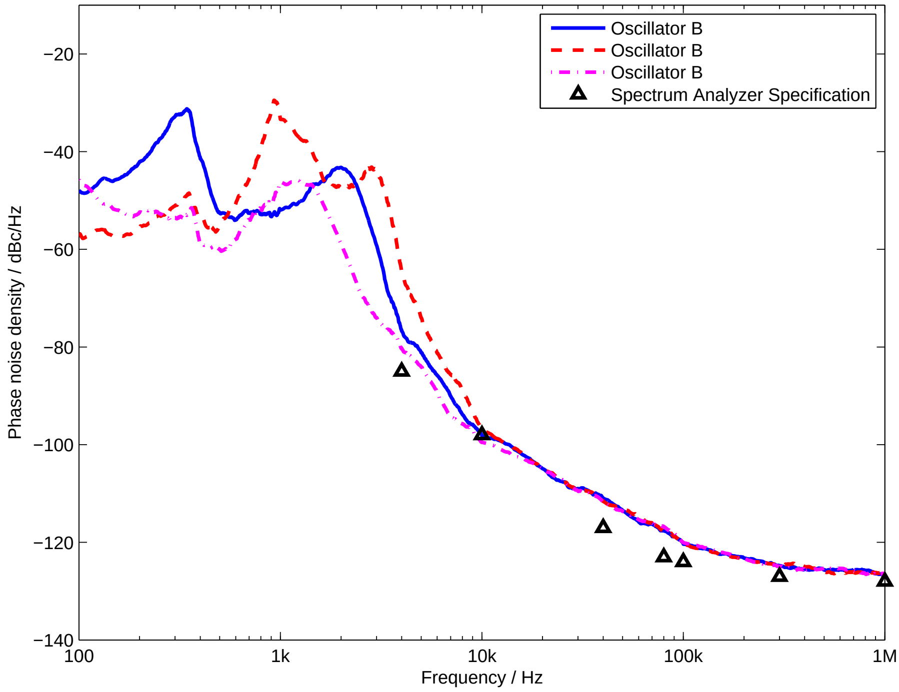


Figure 5.12: Phase noise of the oscillator B measured three times with the E4470B

With 30 to 100 kHz, 4.77 dB is subtracted and with 100 to 150 kHz, 10 dB is subtracted from the results.

Some of the traces in Figure 5.14 start in the larger frequencies, because those oscillators were too noisy, and the measurement could not be locked to the carrier in those low offset frequencies. When the offset frequency is larger, the accuracy for the carrier frequency can be lower and that is the reason why the measurement could be locked with the larger offset frequencies. The accuracy can be lower, because with larger offset frequencies the phase noise changes less with the same frequency change. The level of phase noise falls more rapidly near the carrier than far from it, as Figure 2.2 shows.

5.4 Actual Phase Noise Measurements using the PLL Method

The schematics of the final set-up can be found in Figure 5.15, and Figure 5.16 shows a photograph of the final set-up. There is an optional section in the schematics which can be omitted if the output level of the DUT is in the right region, and if the carrier frequency does not drift too much. The optional attenuator can be used to bring the level near −5 dBm at the input of the mixer. With greater levels the output of the mixer would be distorted. The exact level depends on the mixer used. Too low a signal level would also be a problem. A low noise amplifier can be used to boost the signal from the DUT, if necessary. In addition, if the DUT drifts much, it is necessary to somehow measure the frequency when doing the measurements. It can be done with the optional FFT/spectrum analyzer.

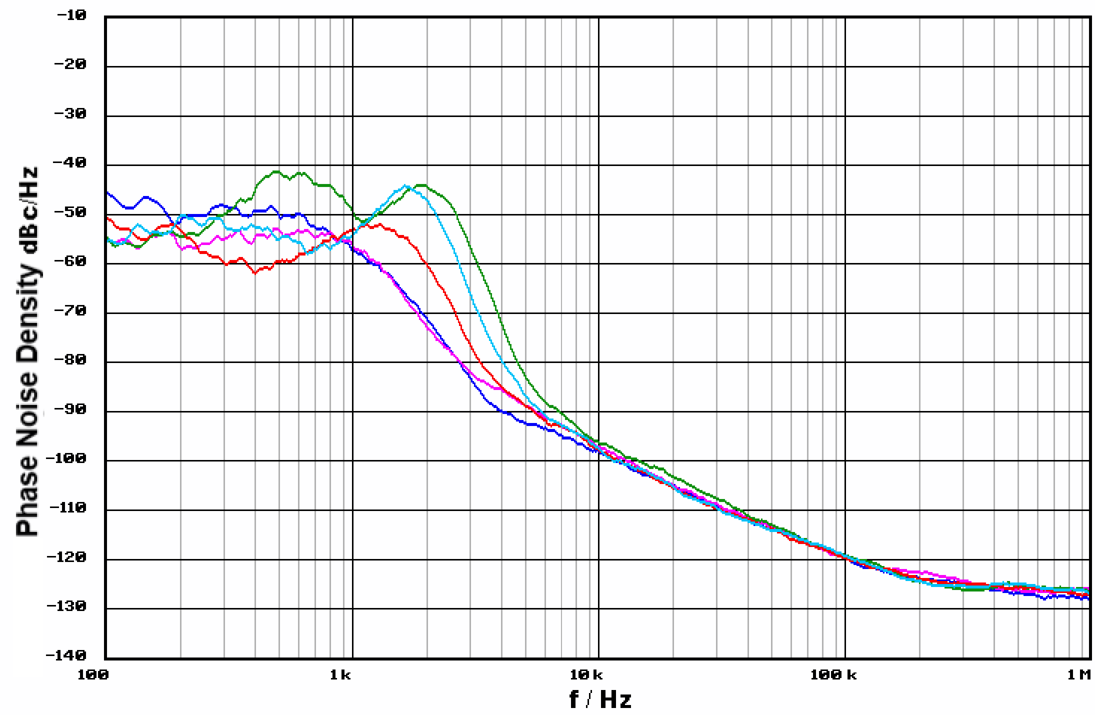


Figure 5.13: Phase noise of the oscillator B measured five times with the KE5FX GPIB toolkit

5.4.1 Measurement Procedure

The actual measurement procedure consists of two different parts. The level of the carrier is measured first. It is done by disconnecting the PLL loop and mis-adjusting the reference oscillator by a few dozens of kilohertz so that there appears a beat note whose frequency is visible in the spectrum analyzer. The frequency of the beat note is the difference between the two oscillators. [22]

In this thesis the frequency of the beat note was usually between 30 kHz and 60 kHz. Figure 5.17 shows the beat note on the screen of the spectrum analyzer. The level of the peak (P_{carrier}) is measured.

It is important to use low enough drive levels so that the beat note is undistorted. When observing the beat note on the display of the spectrum analyzer this can be verified. If the second and third harmonics of the beat note are more than 30 dB lower than the level of the beat note, the signal can be called undistorted. If the harmonics are too high level, the oscilloscope shows a distorted sine wave, so this can also be used as a guide. If the signal is distorted and drive levels can not be lowered, some corrections can be made. These corrections are described in [22] and [7].

After the amplitude of the beat note has been measured, the PLL circuit is connected again and the frequency of the reference oscillator is tuned the required amount to get the PLL locked. When the PLL is locked, there is a DC level shown on the screen of the oscilloscope. The input of the oscilloscope must be set to DC coupling and the triggering

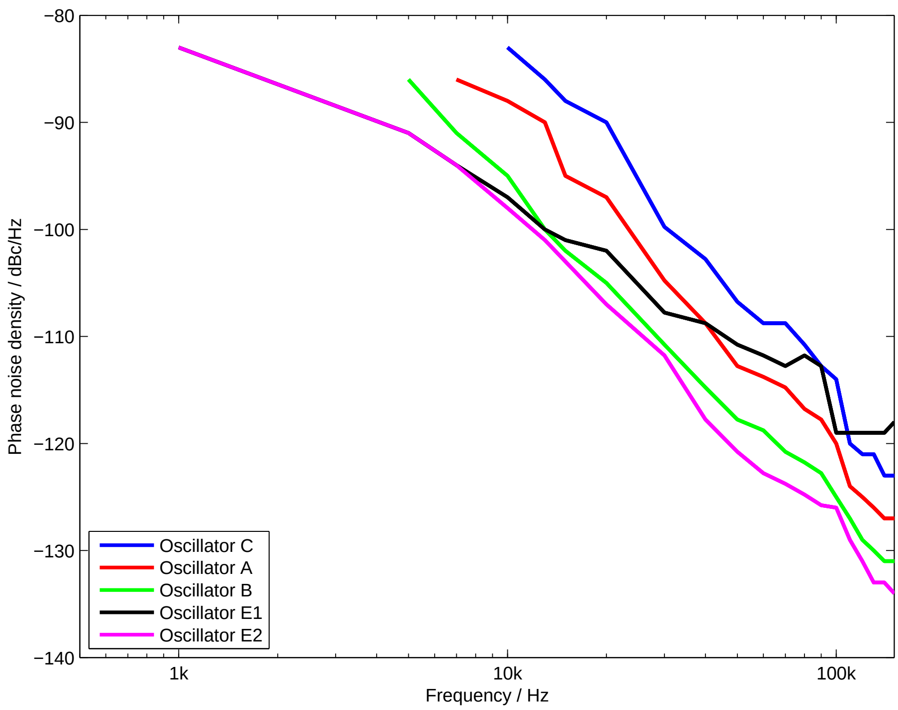


Figure 5.14: Phase noise measured with the spot frequency mode

also for DC. The frequency should be adjusted in a way to get the DC level as close to 0 V as possible. Figure 5.18 shows the error introduced by a DC level far from 0 V. The measured oscillator was E and the used frequency was 422 MHz.

When the PLL is locked, it is time to measure the phase noise levels. Figure 5.17 shows the measured noise level. At this point it is necessary to do some amplitude compensation, to take care of the limited dynamic range of a spectrum analyzer. It can be done either by switching more gain to the amplifier, or less attenuation to the attenuator located just before the spectrum analyzer. The latter one was found easier. When more gain is switched on, PLL usually becomes unlocked, the beat note appears, and the reference oscillator must be adjusted again. At this point it is important to remember that even with 30 dB of gain from the LNAE, the beat note may be of too high level for the spectrum analyzer and thus it is recommended to have some switchable attenuator before the spectrum analyzer. When measuring noisy oscillators, it was sometimes necessary to keep hands on the attenuator and quickly add some attenuation if the PLL became unlocked. If the amplitude of the beat note is high enough, it can break the spectrum analyzer even if attenuation is quickly added. That is why it is important to be very careful during this phase of the measurement. This unlocking can be seen easily and quickly on the oscilloscope, so the oscilloscope should stay connected all the time when doing the measurements.

The effect of different gain settings was also tested, but the results were not promising. It seems to be more convenient to use the same amplifier gain at all times, and just use the switchable attenuator for the amplitude compensation. This way the PLL stays locked

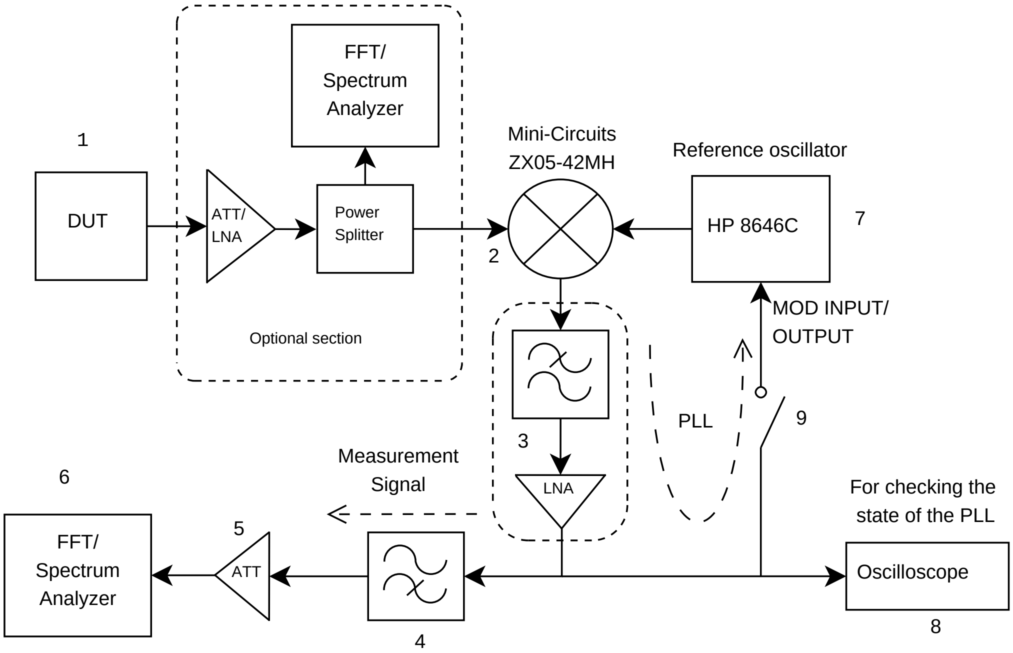


Figure 5.15: The schematics of the final measurement set-up. Numbering is the same in Figure 5.16. The switch 9 is actually the Ext Mod On/Off -button in the spectrum analyzer. It is used to turn the PLL off for the beat note measurement.

better and it is faster and more convenient to do all the measurements. If the amplifier was located outside of the PLL, the gain changes would not alter the PLL and this method could be used better. On the other hand, then it would not be possible to change the loop gain of the PLL and the PLL might not work at all.

5.4.2 Processing the Measurement Results

When the noise level is measured, it has to be processed to get the actual values. The amplitude compensation must be taken care of, values must be transformed to show noise power in a 1 Hz bandwidth, and all additional amplitude related corrections have to be considered. The following equation shows how these different variables affect the result. All variables are in the decibel scale. [5, p. 9.12], [22]

$$\mathcal{L}_{\mathrm{SSB}}\left(\mathrm{dBc/Hz}\right) = P_{\mathrm{N}} - P_{\mathrm{beat}} - G_{\mathrm{comp}} - 10\log_{10}\left(B\right) - G_{\mathrm{correction}} \tag{5.5}$$

where $\mathcal{L}_{\mathrm{SSB}}$ is the actual phase noise density, P_{N} is the measured noise level in dBm, P_{beat} the measured level of the beat signal in dBm, G_{comp} includes all attenuator or amplifier settings in dB, B is the equivalent noise bandwidth of the spectrum analyzer IF filter in Hz, and $G_{\mathrm{correction}}$ includes all other necessary amplitude corrections in dB. These terms will be described in detail in the following text.

Probably the most interesting part in Eq. (5.5) is the $-\log_{10}\left(B\right)$. Because phase noise density is expressed as power in a 1 Hz bandwidth and the measurement usually uses a

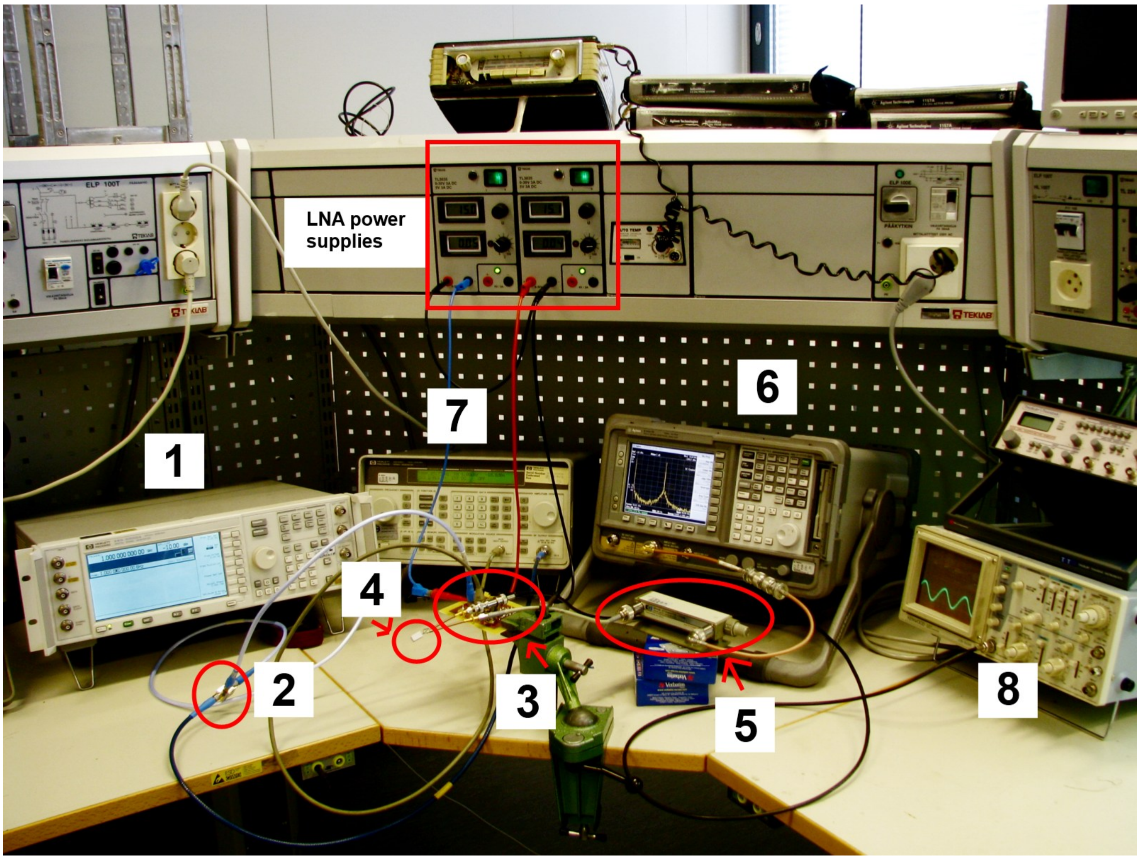


Figure 5.16: The final measurement set-up. The beat note is visible on the spectrum analyzer (6) and on the oscilloscope (8). Numbering is the same in Figure 5.15.

different bandwidth (100 Hz was used in this thesis), the measured level must be transferred to show the actual noise level in the 1 Hz bandwidth. It can be done with the logarithm shown in Eq. (5.5), where B is the equivalent noise bandwidth of the used filter. This is not directly the resolution bandwidth (RBW) of the spectrum analyzer, or frequency resolution of the FFT device, but it must be multiplied with a correction factor. This correction factor depends on the shape of the filter and the result is called equivalent noise bandwidth. Two different filters may have the same 3 dB bandwidths, but different equivalent noise bandwidths.

The correction factor depends on the shape of the filter used. If the shape is known, it can be calculated or found from literature. The shape is not always known well enough. For example Agilent Technologies states that they use 'Gaussian like' or 'near-Gaussian' filters and that the equivalent noise bandwidth is about 1.05 to 1.13 times the 3 dB bandwidth [15]. This depends on the used RBW. The manual of the E4407B spectrum analyzer says that with the RBW in the range of 1 to 300 Hz the filter shape is 'Digital, approximately Gaussian shape' and with the RBW in the range of 1 kHz to 5 MHz it is 'Synchronously tuned four poles, approximately Gaussian shape'. If the correction factor is omitted, there is a small error in the results. With 100 Hz RBW the differ-

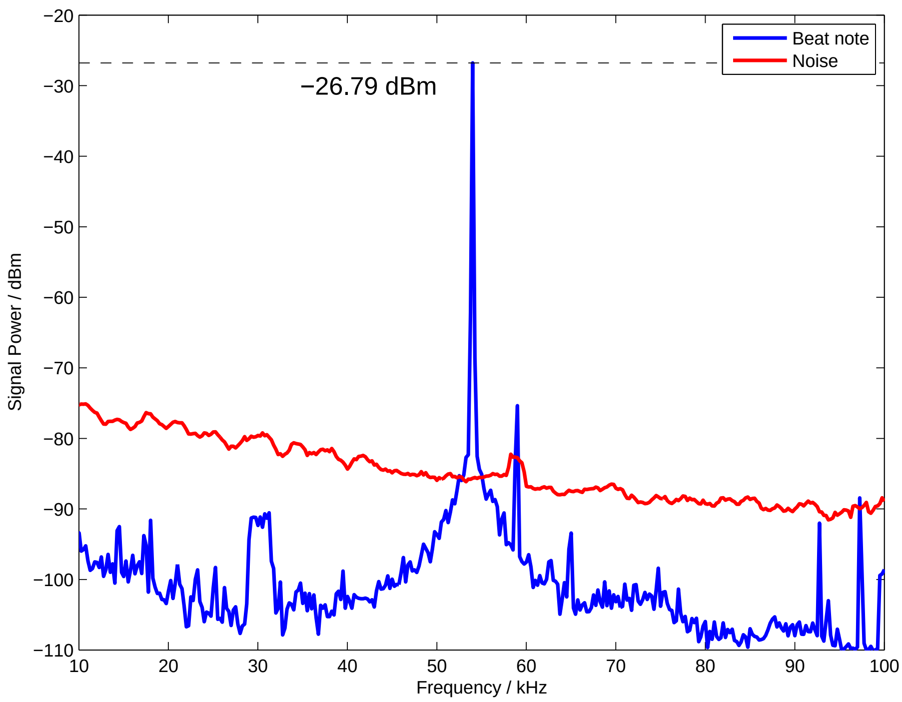


Figure 5.17: The beat note and measured noise level as seen on the screen of the spectrum analyzer

ence in B is 0.53 dB between no correction factor and 1.13 as the correction factor. $\log_{10}(1 \cdot 100\text{ Hz}) = 20$ dB and $\log_{10}(1.13 \cdot 100\text{ Hz}) \approx 20.53$ dB In this thesis a correction factor of 1.1 was used.

In Eq. (5.5) the last part $G_{\text{correction}}$ includes all necessary amplitude correction parameters. Because of the measurement method of most spectrum analyzers, based on the superheterodyne principle, the measured noise level is lower than the actual noise. An instantaneous amplitude of noise has no meaning, because the amplitude of noise is often Gaussian distributed, and can theoretically have any value from zero to infinity. A noise level averaged over some time period is a better way to express the amount of noise. The so called power averaging, which is proportional to RMS voltage, satisfies this requirement. [15]

The Gaussian distributed noise, with standard deviation σ, is band limited as it travels through the IF section of the spectrum analyzer, and it becomes Rayleigh distributed. Video filtering, or averaging is used to get a steady value, the mean value. The Rayleigh distribution has a mean value of 1.253σ. The analyzer is actually a peak-responding voltmeter calibrated to indicate the RMS value of a sine wave. Thus the analyzer must scale its readout by $\frac{1}{\sqrt{2}} = 0.707 = -3$ dB to get RMS value from peak value. Also the mean value of the Rayleigh distributed noise is scaled with the same amount, and that gives us reading that has a mean value of $\frac{1.253\sigma}{\sqrt{2}} \approx 0.886\sigma$ ($-20\log_{10}(0.886) \approx 1.05$ dB below σ). This error is constant, and thus the result can be corrected by adding 1.05 dB to the displayed values. [15]

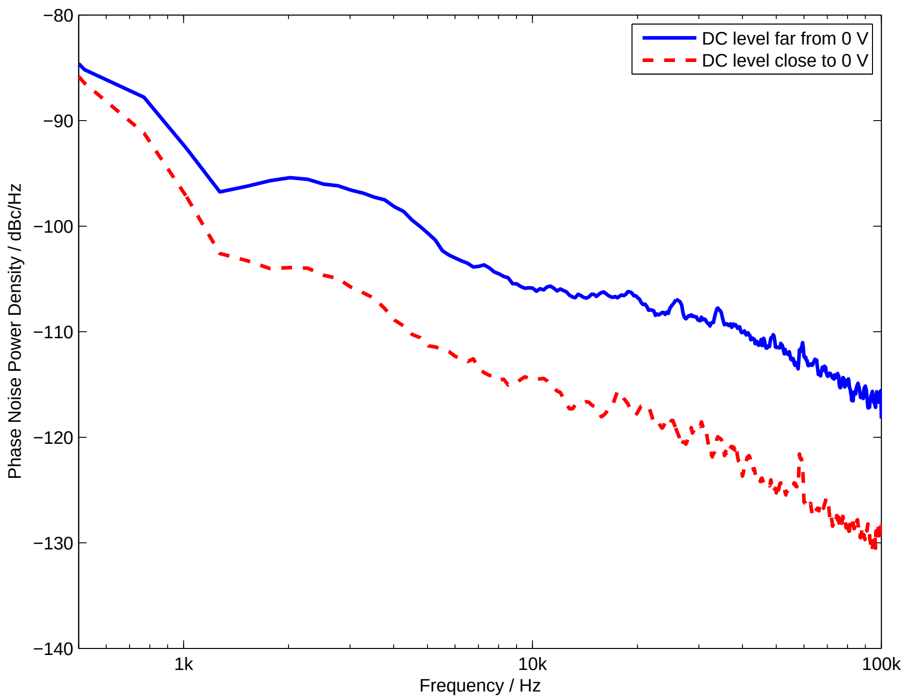


Figure 5.18: Difference in the measured phase noise when using the different DC levels

The display scale of the analyzer is also one error source, but fortunately the error is also a constant and can be corrected. If a logarithmic scale is used, the output of the envelope detector is a skewed Rayleigh distribution. The mean value of this is 1.45 dB lower than the mean value of a non-skewed Rayleigh distribution. Thus the total correction factor is 2.5 dB. The gain of the used logarithmic amplifier is a function of signal amplitude, so lower noise values are amplified more than the higher values. [15]

With an Agilent ESA or PSA Series spectrum analyzer the need for the 2.5 dB correction depends on the used averaging method. There are three different averaging methods: video, voltage, and power. If power averaging is used, no correction is needed because it determines the average RMS level by the square of the magnitude of the signal, not by the envelope of the voltage. In this thesis power averaging was used during the measurements, so the 2.5 dB correction is not needed. [15]

As was stated already, if the phase noise level of the reference oscillator is close to the phase noise level of the DUT, an additional correction factor must be subtracted from the result. This was discussed in Section 3.2. In this thesis, a 3 dB correction factor was used when measuring the oscillator E, and no correction was used for other oscillators. The reference oscillator in all measurements was the Agilent 8648C signal generator. It was assumed that the phase noise of the 8648C and the oscillator E (Agilent ESG-D3000A signal generator) would be similar and thus the correction factor must be used. It was also assumed that the free running oscillators A-C would have worse phase noise performance.

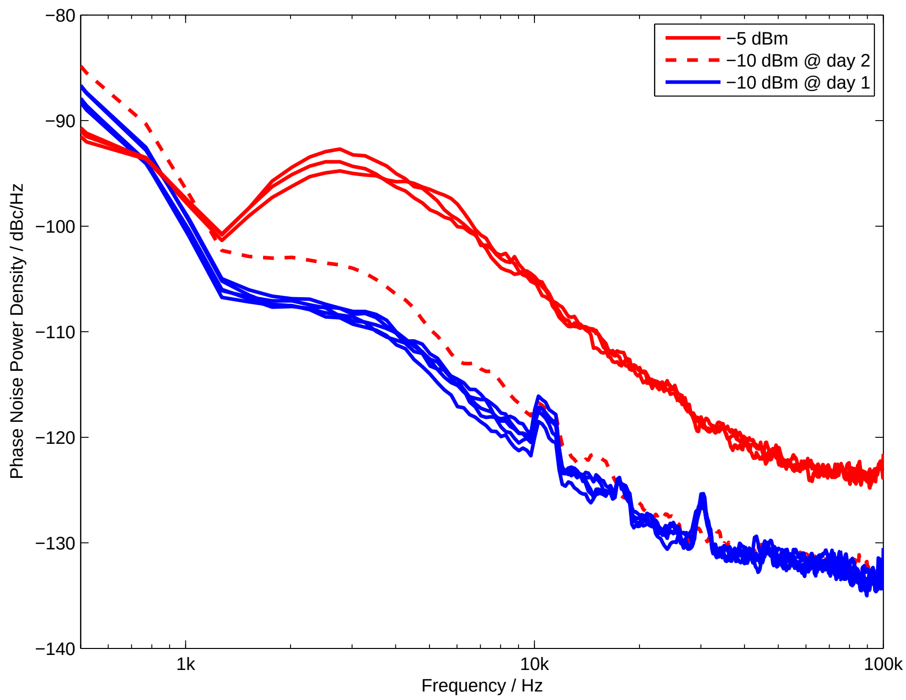


Figure 5.19: Repeatability of the measurement

When mixing signals down to DC, one sideband of the signal is folded onto the other. Because the noise voltages are in phase, they add coherently, and thus additional 6 dB must be subtracted from the measured noise power density. [22] Now the correction factor for the oscillator E can be calculated: $G_{\text{correction}} = 3$ dB+6 dB = 9 dB. For other oscillators $G_{\text{correction}} = 6$ dB.

Figure 5.19 show how well this measurement with the PLL method can be repeated if all the measurement settings and cables used are the same. The same oscillator (E2, frequency 1 GHz) was measured on two different days and naturally the measurement set-up was disassembled and built again in between. On the first day the measurement was repeated five times (traces -10 dBm @ day 1) and the frequency of the reference oscillator was adjusted to get the PLL out of lock between all measurements. The biggest variation is located somewhere around 6 kHz and it is about 2 dB as can be seen from the figure. On the second day it is similar with about 3 dB difference. This difference is probably caused by changes in cabling and adapters, and supply voltage to the LNA. The supply voltage level was not calibrated between the two measurement sessions, but it was set approximately to ± 15 V.

When the output levels of the oscillators were set to -5 dBm, the output is surprisingly different, as can be seen in the third trace group (-5 dBm) in Figure 5.19. This is interesting, because with the higher drive levels the mixer behaved more linearly, if we use the the criterion discussed in the start of this section. With a -10 dBm oscillator power the harmonics were only 29.5 dB lower but with -5 dBm the difference was about 38 dB.

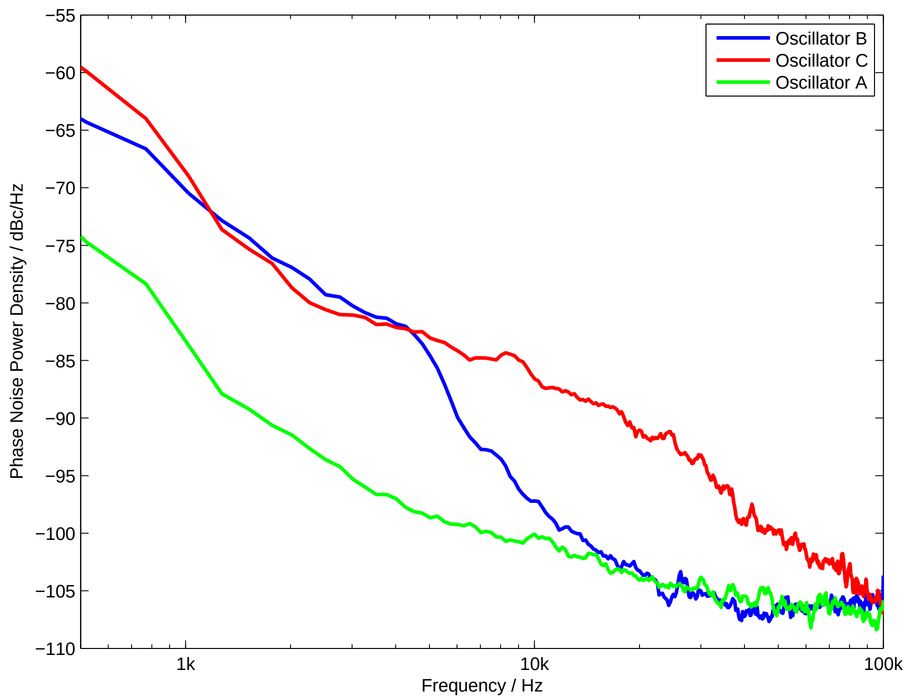


Figure 5.20: Phase noise of the oscillators A, B and C

According to these measurement results, the measured phase noise performance is about 13 dB poorer when using higher drive levels.

Despite this behavior, it can be said that this measurement is reliable and time invariant, when measuring good enough oscillators with the same set-up.

5.4.3 Measuring the free running Oscillators

Finally there are some results from the other oscillators measured during this project. The results can be seen in Figure 5.20. Although it is always interesting to measure devices which are not commercial off-the-shelf products, it was decided to not use those when verifying the performance of the measurement set-up. When verifying something, it is essential to have as few variables as just possible. In this case the amount of variables was made lower by using a good enough oscillator as a DUT. When it was verified that the set-up would work as planned, some real life measurements were made. This kind of oscillators will be measured with this system in the future, so it is a good idea to have also some understanding about that process. It should be noted that the scale of the vertical-axis has been changed, so when comparing these traces to the other traces this should be considered.

It was noticed that even with this kind of a measurement setup, noisy oscillators are problematic. They can be measured, but the results have much variation, and thus it may be necessary to use more statistical processing. The same oscillator should be measured many times, and some statistical methods used to calculate a average phase noise curve.

The statistical methods do not usually take systematic errors in the account, so those must be handled separately. Statistical methods are not covered in this thesis, and it is recommend that the reader should familiarize himself with them if reliable results are wanted.

5.5 Error Analysis

The basis for getting the final phase noise density results with the PLL method (Equation (5.5)) is repeated here for convenience

$$\mathcal{L}_{\mathrm{SSB}}\left(\mathrm{dBc/Hz}\right) = P_{\mathrm{N}} - P_{\mathrm{beat}} - G_{\mathrm{comp}} - 10\log_{10}\left(B\right) - G_{\mathrm{correction}}$$

The exact error limits in all terms are unknown, but this section will give some insight in the amount of deviation in the results that can be expected.

It was noticed during the measurements that even when all the settings should be similar, there can be 3 dB difference between the measured noise levels. So, it can be assumed that the variation in P_{N} is ± 1.5 dB. There are some areas in the measurement set-up which cause this variation. First of all, even if the same cables and adapter are used all the time, there are some variation in how well the connectors are mated and how cables are routed. Also a small difference in the supply voltage of the low noise amplifier can cause some variation, as already discussed. The temperature around the measurement set-up will also have an effect.

The variation in P_{beat} (the measured beat note power) depends mostly on the spectrum analyzer, and can be assumed small, around ± 0.25 dB. The accuracy of G_{comp} (the amplitude compensation) depends on the method used to measure all the attenuators and LNA. ± 0.25 dB is a realistic assumption for this term, because the measured attenuation values were compared to the calibrated values and they were inside the ± 0.2 dB range.

The biggest variation in $10\log_{10}(B)$-term is caused by the shape factor of the applied filter. Some discussion can be found at Section 5.4.2. B is calculated as $B = \mathrm{RBW} \cdot S$. With the resolution bandwidth (RBW) set to 100 Hz, the difference in $-10\log_{10}(B)$ term between the shape factor $S = 1$ and $S = 1.13$ is 0.53 dB. So, a realistic variation is ± 0.25 dB.

For the last term, $G_{\mathrm{correction}}$ (includes all additional corrections), a realistic assumption for the variation is ± 0.5 dB. The variation in the last term is caused by problems in determining how close the phase noise performances of the DUT and the reference oscillator are to each other.

Experimentally it was seen that a ± 2 dB variation in phase noise density is a realistic expectation for the final results. This can be seen in Figure 5.19. Some more in depth error analysis can be made with the root-sum-square (RSS) uncertainty or the maximum uncertainty methods. Even though these methods are not designed for handling numbers

Table 5.3: The uncertainty results

Term	Value	Uncertainty
P_{N}	$-80\ldots-60$ dBm	±1.5
P_{beat}	$20\ldots40$ dB	±0.25
G_{comp}	$-30\ldots-20$ dB	±0.25
$10\log_{10}(B)$	$20\ldots21$ dB	±0.25
$G_{\mathrm{correction}}$	$6\ldots11.5$ dB	±0.5
RSS uncertainty		±1.6 dB
Maximum uncertainty		±2.8 dB

in the decibel scale, it is a practice in the RF field to use them for that. Using numbers in the decibel scale introduces some error, but with small numbers the error is also small, so it can be neglected.

Reference [25, p. 19] defines the RSS error as

$$E = \sqrt{\sum_{i=1}^{n}\left(\frac{\partial f}{\partial X_i}E_{\bar{X}_i}\right)^2} \tag{5.6}$$

where f is the function under investigations, here the sum, X_i is a parameter of f, and $E_{\bar{X}_i}$ is the estimated error in term X_i. The maximum error occurs when the maximum error is realized in all the terms and all the errors are in the same direction [25, p. 19]:

$$E_{\max} = \sum_{i=1}^{n}\left|\frac{\partial f}{\partial X_i}\right|E_{\bar{X}_{i,max}} \tag{5.7}$$

So, the RSS error estimate for $\mathcal{L}_{\mathrm{SSB}}$ is

$$\begin{aligned} E_{\mathcal{L}_{\mathrm{SSB}}} &= \sqrt{E^2_{P_{\mathrm{N}}} + E^2_{P_{\mathrm{beat}}} + E^2_{G_{\mathrm{comp}}} + E^2_{10\log_{10}(\mathrm{B})} + E^2_{G_{\mathrm{corr}}}} \\ &= \sqrt{1.5^2 + 0.25^2 + 0.25^2 + 0.25^2 + 0.5^2} = \sqrt{2.6875} \approx 1.64\,(\mathrm{dB}) \end{aligned} \tag{5.8}$$

The ±1.64 dB error is quite close to the ±2 dB error that was estimated experimentally, so it is quite safe assumption. The maximum error in this measurement is:

$$\begin{aligned} E_{\mathcal{L}_{\mathrm{SSB}},\max} &= E_{P_{\mathrm{N}}} + E_{P_{\mathrm{beat}}} + E_{G_{\mathrm{comp}}} + E_{10\log_{10}(\mathrm{B})} + E_{G_{\mathrm{corr}}} \\ &= 1.5 + 0.25 + 0.25 + 0.25 + 0.5 = 2.75\,(\mathrm{dB}) \end{aligned} \tag{5.9}$$

Table 5.3 gathers together the uncertainty results with the information about the order of magnitude of all the terms.

6. CONCLUSIONS

In this chapter the whole thesis is summarized. Conclusions are drawn from the different parts and the thesis in general. The goals of the thesis are discussed and the results covered. There are also some suggestions what can be done to make the system better, and what things could not be covered in this thesis but should be investigated and verified in the future.

6.1 Goals of the Thesis

The goals of this thesis were to investigate phase noise and different measurement systems, and to build and verify the performance of one phase noise measurement system that could be used in the RF laboratory. These goals were fulfilled, although some things were left open for future research.

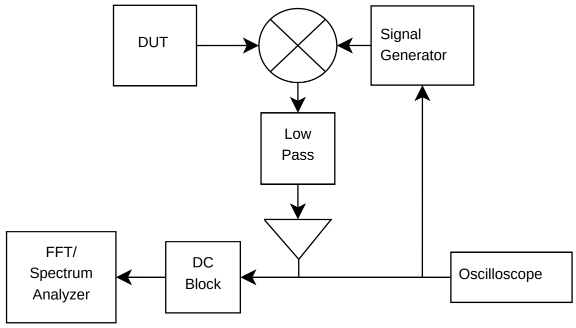
A usable measurement system resulted from this project. It is still an experiment version without any casing, for example. It was verified that the system could be used for phase noise measurements, and that was one of the goals. It can thus be said that the project was successful.

6.2 Future Research

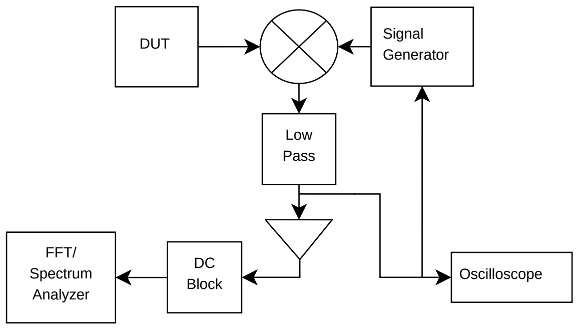
The noise properties of the low noise amplifier were not verified. Experimentally it was seen a low noise one when using the 30 dB gain settings, but no measurement data was gathered. Its performance should thus be verified and probably some fine tunings could be made to make it even less noisy.

The build quality and form factor of the amplifier could be improved. At the end of this project it is just a plain PCB board without any case. A metal case with real connectors, especially for the supply voltage, should be considered. It would also be wise to include to the board the now external DC block, which protects the spectrum analyzer, because then two connectors could be excluded from the system, and probably the noise properties would also improve a little. In this context build quality means that the manufactured amplifier was just a test version built to verify the performance of this method. It can be used for actual measurements, but it is not made to withstand the usual student laboratory handling.

It must also be said that the current PCB layout has some problems. It was probably the tenth version, and was altered dramatically just before the last version which was used for



(a) Set-up used in this thesis



(b) Alternative set-up

Figure 6.1: Two variations of the same measurement set-up

the actual measurements. There is for example not enough space for the supply voltage connectors. To reduce the required amount of adapters, the other output BNC connector should be changed to an SMA connector, and it would probably be wise to include a third output connector.

An even better improvement would be to include also the mixer to the system. Right now it was time consuming to build and dismantle the measurement setup. It took usually about 15 to 20 minutes to assemble the system, because many adapters and different cables had to be used. A better system would have inputs from the DUT and the reference oscillator and outputs for the spectrum analyzer, PLL, and the oscilloscope. If the connectors are chosen properly there will be no need for adapters. With careful planning and design even the switchable attenuator could be included in this device. It would make the measurement process much easier and results more repeatable, because fewer connections would have to be made.

One other interesting variation is to get signal to the PLL and oscilloscope right after the mixer and low pass filter. The difference is shown in Figure 6.1. The amplifier would amplify only the signal going to the spectrum analyzer. By doing it this way, the amplitude compensation during the measurement process could be more easily made with the gain switch of the amplifier. The down side is that then it would be impossible to tune the loop gain of the PLL. It should naturally be verified, whether this would work at all, because the loop gain could not be changed.

It would be also interesting to study whether a noisy VCO could be measured better with the set-up used in this thesis, or with the alternative setup where the reference oscillator is the free running one, and the DUT is included in the PLL loop. This could not be tested during this process, because a suitable PLL amplifier was not available and the tight schedule did not allow to build one. A suitable PLL amplifier takes the output of the mixer ($\pm$ some dozens of millivolts) and generates a suitable tuning voltage for the VCO. The suitable tuning voltage is usually 0 to 5 V DC, but a good PLL amplifier would have an adjustable tuning voltage.

Even though there are some topics left for future research, it can be said that the measurement system built during this project can be used to measure phase noise of oscillators. Measuring phase noise of low noise oscillators is not an easy task, but it can be done without expensive commercial phase noise measurement systems using only the equipment available in most RF laboratories.

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A. LNA SCHEMATICS AND PCB LAYOUT

This appendix includes the schematics and the PCB layout of the low noise amplifier designed for the PLL method measurement set-up. The sizes of the etching masks are correct, so they can be directly copied if the reader wants to build the same amplifier and has the same parts available. The PCB was designed with Eagle Layout Editor [26], and the Eagle files of this amplifier can be obtained by sending an email to olli.rajala@ravoltek.net.

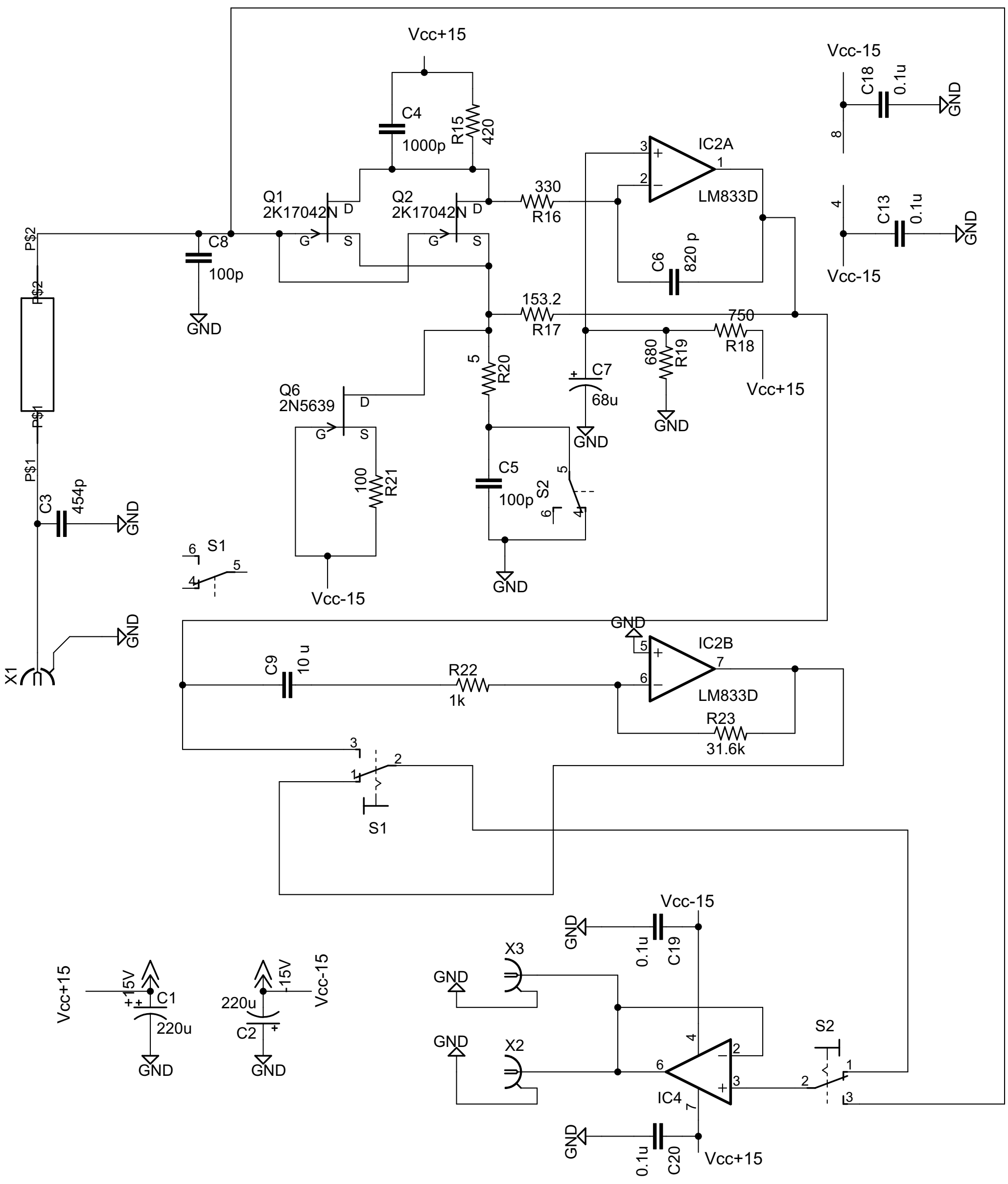
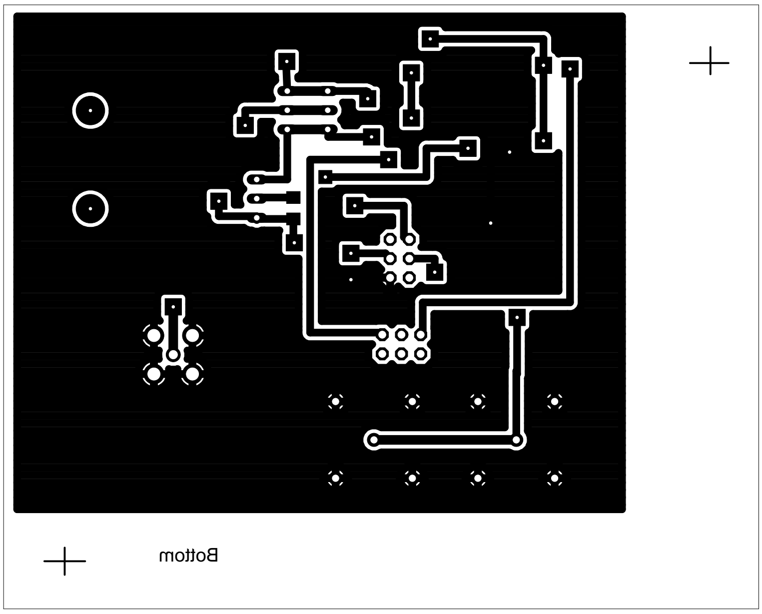


Figure A.1: The schematics of the low noise amplifier

Table A.1: LNA parts

Part	Value	Package
C1	220u	SMC_D
C2	220u	SMC_D
C3	454p	C1206
C4	1000p	C1206
C5	100p	C0805K
C6	820 p	C1206
C7	68u	E/7260-38R
C8	454p	C1206
C9	10 u	C1206
C13	0.1u	C1206
C18	0.1u	C1206
C19	0.1u	C1206
C20	0.1u	C1206
IC2	LM833D	SO08
IC4	AD825	SO08
Q1	2K17042N	TO92-
Q2	2K17042N	TO92-
Q6	2N5639	TO92-
R15	420	R1206
R16	330	R1206
R17	153.2	R1206
R18	750	R1206
R19	680	R1206
R20	5	R1206
R21	100	R1206
R22	1k	R1206
R23	31.6k	R1206
S1	Omron A9T	A9T21-0011
S2	Omron A9T	A9T21-0011
U$1	2.2u	own inductor
X1		BU-SMA-V
X2		A1944
X3		A1944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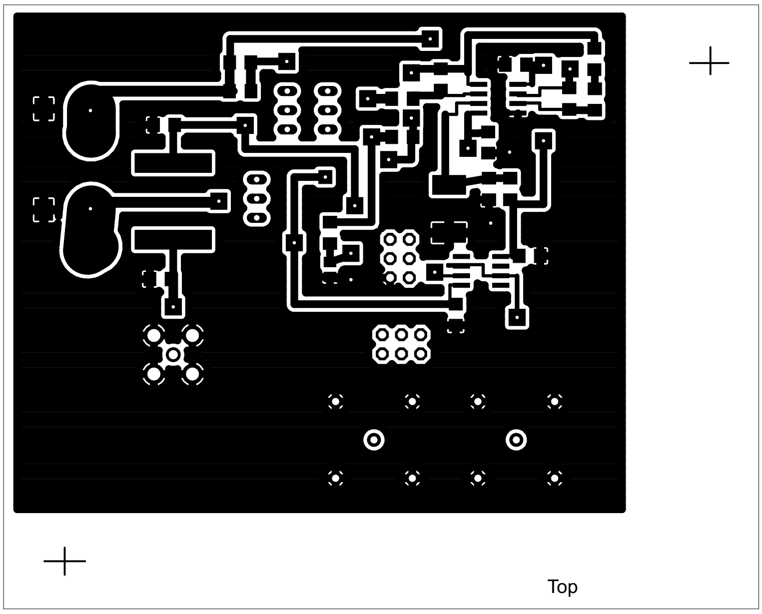
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